Customer No. 22,852
Attorney Docket No. 08364.0062-01000

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application of:	)
Aled Wynne JONES et al.	) Group Art Unit: 2857
Application No.: 10/777,127	) Examiner:
Filed: February 13, 2004	)
For: TAG TRACKING	)

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

Sir:

CLAIM FOR PRIORITY

Under the provisions of 35 U.S.C. § 119, Applicants hereby claim the benefit of the filing date of Great Britain Patent Application Nos. 0119787.0, filed August 14, 2001; 0206597.7, filed March 20, 2002; 0209781.4, filed April 29, 2002; 0027886.1, filed November 15, 2000; and 0027888.7, filed November 15, 2000, for the above-identified U.S. patent application.

In support of this claim for priority, enclosed is one certified copy of each priority application.

Respectfully submitted,

FINNEGAN, HENDERSON, FARABOW, GARRETT & DUNNER, L.L.P.

Dated: August 16, 2004

By: ____________________
Ernest F. Chapman
Reg. No. 25,961

EFC/FPD/gah
Enclosures

The Patent Office

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The Patent Office
Concept House
Cardiff Road
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South Wales
NP10 8QQ

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I, the undersigned, being an officer duly authorised in accordance with Section 74(1) and (4) of the Deregulation & Contracting Out Act 1994, to sign and issue certificates on behalf of the Comptroller-General, hereby certify that annexed hereto is a true copy of the documents as originally filed in connection with the patent application identified therein.

In accordance with the Patents (Companies Re-registration) Rules 1982, if a company named in this certificate and any accompanying documents has re-registered under the Companies Act 1980 with the same name as that with which it was registered immediately before re-registration save for the substitution as, or inclusion as, the last part of the name of the words "public limited company" or their equivalents in Welsh, references to the name of the company in this certificate and any accompanying documents shall be treated as references to the name with which it is so re-registered.

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Signed M Cooki

Dated 29 March 2004

THE PATENT OFFICE
G
20 MAR 2002
LONDON

21MAR02 E705246-1 D02917
P01/7700 0.00-0206597.7

Patents Form 1/77
Patents Act 1977
(Rule 16)

The Patent Office

Request for grant of a patent

The Patent Office
Cardiff Road
Newport
South Wales NP10 8QQ

1. Your reference
1874701/AM

2. Patent Application Number **0206597.7** 20 MAR 2002

3. Full name, address and postcode of the or of each applicant *(underline all surnames)*

Scientific Generics Limited
Harston Mill
Harston
Cambridgeshire
CB2 5GG

Patents ADP number *(if known)* 7970296002

If the applicant is a corporate body, give the country/state of its incorporation
Country: **England**
State:

4. Title of the invention

PHASE BASED RF LOCATION SYSTEM

5. Name of agent **Beresford & Co**

"Address for Service" in the United Kingdom to which all correspondence should be sent
2/5 Warwick Court
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London WC1R 5DH

Patents ADP number 1826001

6. Priority details

Country	Priority application number	Date of filing

Patents Form 1/77

7. If this application is divided or otherwise derived from an earlier UK application give details

Number of earlier application | Date of filing

8. Is a statement of inventorship and or right to grant of a patent required in support of this request?

Yes

9. Enter the number of sheets for any of the following items you are filing with this form.

0	Continuation sheets of this form
58	Description
0	Claim(s)
0	Abstract
0	Drawing(s)

10. If you are also filing any of the following, state how many against each item.

0	Priority documents
0	Translations of priority documents
1 + 5 copies	Statement of inventorship and right to grant of a patent *(Patents form 7/77)*
0	Request for preliminary examination and search *(Patents Form 9/77)*
0	Request for Substantive Examination *(Patents Form 10/77)*
0	Any other documents *(please specify)*

11. I/We request the grant of a patent on the basis of this application

Signature Beresford & Co Date 20 March 2002

BERESFORD & Co

12. Name and daytime telephone number of person to contact in the United Kingdom

Alan MacDougall

Tel: 020 7831 2290

Patents Form 7/77
Patents Act 1977
(Rule 15)

The Patent Office

Statement of inventorship and of right to grant of a patent

The Patent Office
Cardiff Road
Newport
South Wales NP10 8QQ

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1. Your reference
1874701/AM

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2. Patent Application Number
accompanying application reference 1874701

0206597.7

20 MAR 2002

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3. Full name of the or each applicant

Scientific Generics Limited

---

4. Title of the invention

PHASE BASED RF LOCATION SYSTEM

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5. State how the applicant(s) derived the right from the inventor(s) to be granted a patent

By virtue of employment.

---

6. How many, if any additional Patents Forms 7/77 are attached to this form?

One

---

7. I/We believe that the person(s) named over the page (and on any extra copies of this form) is/are the inventor(s) of the invention which the above patent application relates to.

Signature Beresford & Co Date 20 March 2002

BERESFORD & Co

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8. Name and daytime telephone number of person to contact in the United Kingdom

Alan MacDougall

Tel: 020 7831 2290

SUPPLEMENTARY SHEET FORM 7/77

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Phase Based RF location System

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CONTENTS

1 DOCUMENT INTRODUCTION

This document contains all the technical file-notes used to support the feasibility of a personnel tracking system in an urban environment.

- Section 2 describes two phase based location systems and leads onto a new, enhanced system that combines the two to give new benefits and capabilities.
- Section 3 describes the propagation and pathloss calculations and results used for coverage modelling. These cover the Licence exempt bands below 3GHz. It also covers base station density calculations for FDMA and TDMA based systems.
- Section 4 describes the urban RF propagation coverage model used to indicate the number of base stations needed to cover a typical village area, using 433, 868 and 2400MHz bands.
- Section 5 describes the multipath environment for both indoor and outdoor scenarios, providing typical figures to consider.
- Section 6 discusses various multipath mitigation techniques considered for protecting a phase based location system robust in the environment outlined in Section 5.
- Section 7 describes the improvement having redundant base stations has on location measurements based on multipath affected signals.
- Section 8 describes the multipath propagation model developed and used to assess the effect multipath has on a phase based location system in a user defined environment. Several multipath mitigation techniques are also modelled and their effectiveness demonstrated.

2 ENHANCED PHASE BASED LOCATION SYSTEM

The system proposed combines a phase difference location system between base stations and a phase difference system between signals from the same base station. This section, will cover the following:

- Summary of two systems using the individual techniques, in general terms,
- How we propose to combine them and the benefits
- Example combined system
- Example system accuracy and outline DSP requirements for the receiver
- Base station identification and calibration

2.1 Summary of Core systems

2.1.1 Transmitter phase difference location system

Figure 1 shows the general architecture of this system. It was originally covered by patent US3889264 but is now in the public domain. This system uses multiple fixed base stations transmitting similar carrier frequencies to a mobile receiver. The receiver measures the phase difference directly between the received carriers and uses this plus the known location of the base stations to calculate a fix on its location using a hyperbolic algorithm. A minimum of 3 transmitters are required if they are synchronised, 4 otherwise, extra ones provide surplus information to enable minimisation techniques to be used to improve reliability. There is a cyclic ambiguity problem due to the wavelength of the carrier repeating and producing multiple solutions. This is usually solved with a seed location for the algorithms, typically the initial location or region the receiver is in. If the transmitters aren't synchronised, further clock variations and initial phase differences of the transmitters can be calibrated out with a reference receiver located at a known location. The calibration information is then sent to the mobile receiver.

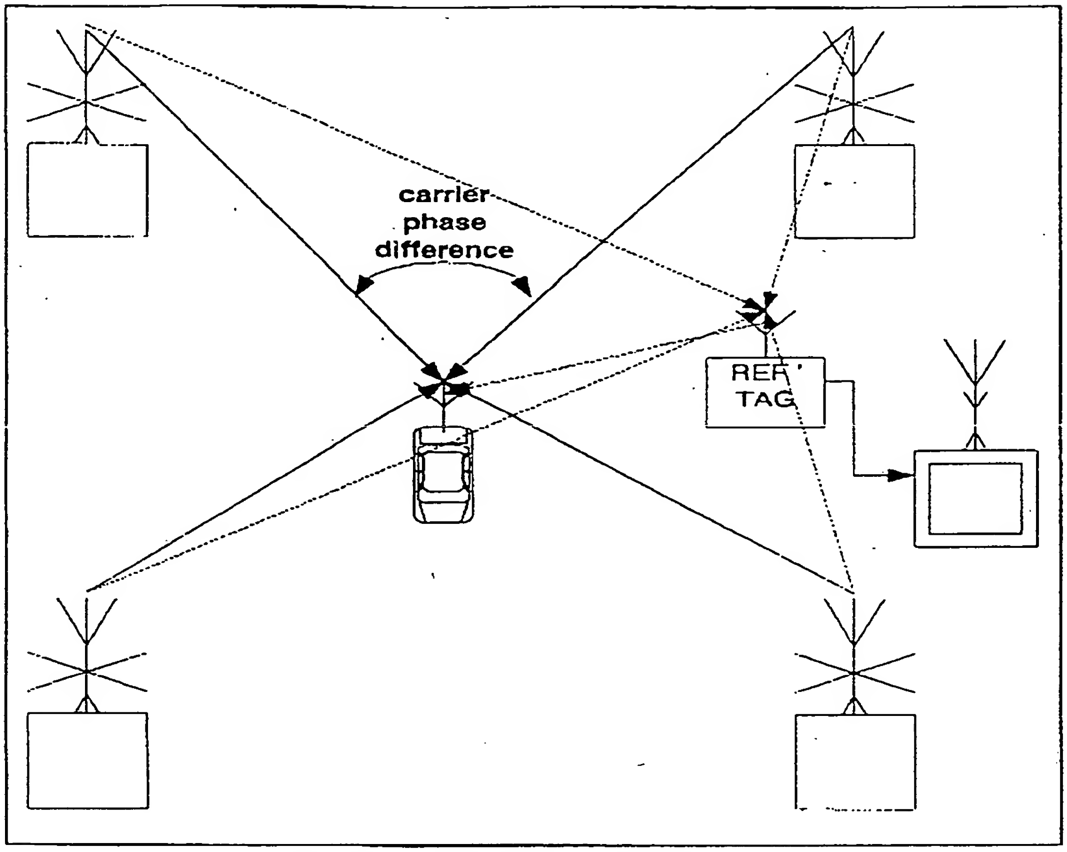


Figure 1: NASA IP based location system

2.1.2 Multiple tone phase difference location system

See Figure 2 for the overall system architecture for the inverted Racetrace solution (the multiple tone phase difference system). The original Racetrace system was designed for dumb mobile tag transmitters and an intelligent receiver system, so the system calculated and knew the location of the tags rather than the tags knowing where they were. An alternative system is to invert this, to enable the tag to hold the location information. The system will therefore consist of multiple fixed base stations transmitting multiple carrier tones at defined frequency separations, to a mobile receiver. The receiver measures the phase difference between tones from the same base station ascertaining distances from them and hence location. If the system isn't synchronised, further clock variations and initial phase differences of the transmitters can be calibrated out with a reference receiver located at a known location.

Like the previous system (Section 2.1.1), there is a cyclic ambiguity problem as the possible location solutions repeat themselves depending on the wavelength of the frequency differences being measured. This can be resolved by forcing a

boundary condition for the initial location solution to be in. Further, by nesting the frequency pairs so the separations increase, a coarse solution can be calculated from two close tones and used as a seed for a solution from a wider pair of tones, to remove the ambiguity for them. As such, the system is totally scalable according to the frequency separations used. The tones could equally as well be PN coded signals or any other signal with defined, repeatable phase characteristics.

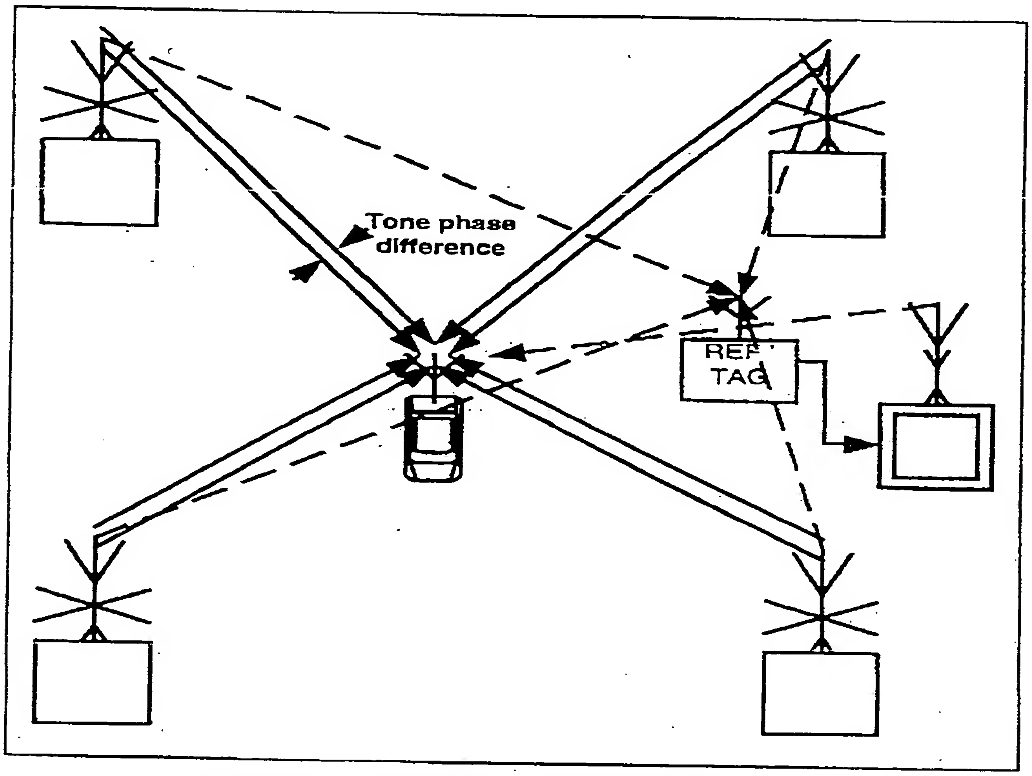


Figure 2: (Inverted) RaceTrace based location system

The actual current implementation uses individual tones that are transmitted sequentially rather than simultaneously, so there is a time/phase offset that must be extrapolated out. All tones are compared at a common time for the phase difference to maintain accuracy. Any sequential phase system will encounter this issue. If base stations transmit at different times with respect to each other as in TDMA, a similar time re-referencing process will be needed on the individual base station results to give the final position.

2.2 Proposed new combined system

2.2.1 Benefits of a combined system

It is proposed to combine the two phase based systems to achieve the following benefits:

- New protected IP (already filed)
- Selective / weighted combining of each systems' results
- Use of one system to seed the other to remove cyclic ambiguity
- Facilitates cross band / system operation which offers different multipath channels
- Different system propagation path dependencies. The phase difference are measured by multi-tone/BS and 1 tone from several BS.
- Totally scalable system, only dependant on permitted RF bandwidths
- Only partially RF carrier independent. Balance between systems and accuracy requirement decides.

Summary description

The new system consists of a multitude of base stations at known locations, each transmitting tones of specific frequency separations to one or more mobile receivers at unknown locations. The receivers measure the phase of all incoming tones and

a) Calculate the phase difference between different tones from the same base station

b) Calculate the phase difference between similar tones from different base stations.

This information, combined with the base station locations, forms the input to the location algorithm and minimisation to provide the position result. The system is initially applied to radio frequency carriers, but could be applied to acoustic or optical as well. Only the implementation techniques would change.

Although the above is what will be described for the rest of the document, depending on the application requirements, the whole of the system could be inverted so the base station side of the system performed the location calculations and knew the position, instead of the mobile. A mobile transmitter could be tracked by base station receivers that measured tone phase differences between stations as well as tone pair differences at each base station. The same core theory and benefits apply.

The tones mentioned in the document could equally as well be any other signal with defined, periodic phase characteristics. This includes RF carriers, signals

modulated with PN codes (whose code length and chip-rate define their equivalent wavelength for use in the location system) or other information signals.

Phase difference between different tones from the same base station.

Through careful selection of frequency differences, an appropriate distance scale may be generated based on the wavelength of that frequency difference. The distance from a base station is based on the phase difference and compared to that scale.

i.e. 2π phase difference $=n\lambda$

$\pi/2$ phase difference $=n\lambda+\lambda/4$

Figure 3 shows how tone phase difference equates to distance and also the cyclic ambiguity represented by n above. With phase information from several base station signals, a hyperbolic algorithm provides possible positions for the mobile position with respect to the base stations.

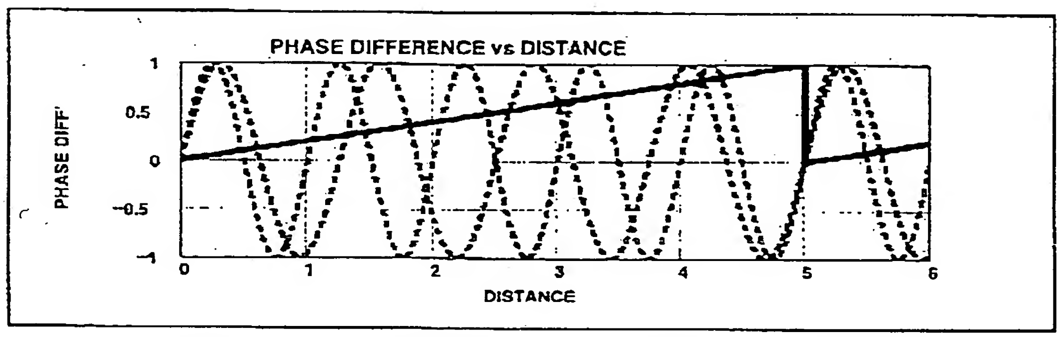


Figure 3: Representation of phase difference vs distance

There is a further unknown phase parameter to be identified before the final distance can be actually calculated. This is the initial phase between the two tones at the point of radiation from the transmitter at that instant in time. This can be calibrated out with synchronised transmissions of a reference receiver placed at a known location picking up the same signals and sending the converted calibration data to the network of direct to the mobiles.

The single correct location can be selected if the cyclic ambiguity is removed from the calculations. This is typically either through seeding with information of which cycle the position is in, or through application of boundary restrictions. If a boundary restriction is set around the whole zone, a pair of tones for coarse measurement covering the whole zone within a single cycle can provide a seed for finer resolution tones. Hence by nesting the tones, a fine resolution position may be calculated over an arbitrary zone size.

Phase difference between similar tones from different base stations

As with the above scheme, careful selection of the frequency yields a distance scale based on the wavelength, comparable to the zone to be covered. The position from the base stations is based on the phase difference between the carrier received from each station and compared to that scale. The phase to distance conversion calculations are similar to those mentioned above and in Figure 3. The cyclic ambiguity and initial transmit phase unknown can also be dealt with in the same way. The key difference is that the wavelength scale is measured between the stations rather than to them.

For example, if two base stations are placed a wavelength apart and the carriers were sent with the same initial phase, a zero phase difference would be measured anywhere on the line directly between them. A phase difference of 90^0 would be measured at a point ¾ of the way directly between them and on all points in an arc represented by the loci of the solution to the equations. To locate the exact position on any of these loci, a third base station is needed to provide another set of loci. The location will be given by where they cross. On larger scales there is a cyclic ambiguity problem.

Combined system

As the initial signal measurement required are the same for both systems, the hardware and front end processing is also the same. By combining the location results from the two systems, with quality measure weightings, enhanced accuracy can be obtained. This also leads to increased flexibility in system design when looking at frequency choice, bandwidths and propagation environments.

2.3 Example combined system

For a 1km zone area with a 0.5m accuracy requirement, what could a system look like if only Licence exempt bands were used?

Table 1 shows the key characteristics of the European licence exempt bands below 3GHz

BS-BS measurement only

If a phase measurement accuracy of 1% were assumed for all phase measurements, a measurement on a 6.7MHz signal would give a 44m scale and 0.44m resolution. A similar measurement at 27MHz would give 11cm accuracy.

BS tone-tone measurement only

A measurement of two tones within 200KHz of the 27MHz band would give a 1.5Km scale and 15m resolution. A similar measurement made cross-band, between 6.7MHz and 13.5MHz would give a 44m scale and 0.44m accuracy. A further measurement cross-band between 13.5MHz and 27MHz would give 22m scale and 0.22m accuracy.

Combined system

By careful selection of the combinations, and coarse calculations being used to seed fine accuracy calculations, cyclic ambiguity can be avoided and a scalable system from Kms to cms can be achieved.

- The 1.5Km BS tone-tone scale would cover the zone (200KHz bandwidth).
- The 15m BS tone-tone accuracy would provide a seed for both of the 44m scale finer resolution calculations giving 0.44m accuracy (6.7MHz bandwidth and 6.7MHz carrier)
- Both 44m scale calculations provide 0.44m accuracy and so can be weighted and combined intelligently or just averaged, depending on desired algorithm complexity (this will reduce the multipath problem as the different solutions will be affected by different multipath components).
- Further accuracy can be achieved by using the 22m scale to get to 0.22m accuracy (the 13.5MHz bandwidth and the 13.5MHz carrier) or the 11m scale to get to 0.11m accuracy (the 27MHz carrier) using the 44m scale as a seed.

Other factors affecting the measurement accuracy also need to be considered when selecting frequencies and frequency differences, e.g. multipath, propagation loss etc, but the core process remains.

Frequency (MHz)	6.765-6.795	13.553-13.567	26.957-27.283	40.66-40.7	433.05-434.79	434.04-434.79	868-868.6	868.7-869.2	869.3-869.4	869.4-869.65	869.7-870	2400-2483.5
BW (MHz)	0.03	0.014	0.326	0.04	1.74	0.75	0.6	0.5	0.1	0.25	0.3	83.5
Power	42dBuA/m @10m		10mW	10mW	a)10mW b) 1mW, 13dBm/10KHz	10	25mW	25mW	10mW	500mW	5mW	a)10mW b)100mW
Duty cycle	-	-	10%	-	a)10% b)100%	100	1%	0.1%	-	10%	100%	-
Channel spacing	-	-	-	-	-	25KHz	-	-	25KHz	25KHz	-	-
Carrier λ	44m	22m	11m	7.4m	0.7m	0.7m	0.35m	0.35m	0.35m	0.35m	0.35m	0.12m
Frequency difference λ	10Km	21km	920m	7.5Km	170m	400m	500m	600m	3Km	1.2Km	1Km	3.5m

Table 1: European licence exempt bands under CEPT-REC-70-03

2.4 Example measurement accuracy & dsp

- For 1% rms total positioning accuracy, we need to measure phase to 3.6 degrees, i.e. +/- 1.8 degrees rms.
- If we take the existing Racetrace S/N requirement of 20dB as the best conditions we can expect for a single measurement, (ignoring multipath) this equates to 5.7 degress rms.
- By taking "n" readings, we can reduce the rms error by $\sqrt{n}$, so the number of measurements needed to reach the target is:-

  $(5.7/1.8)^2 = n = 10$
- If the signals and channel filtering are processed in DSP, and an FFT used for the processing, the FFT buckets need to be narrow enough to isolate the wanted signals from possible interferers. 500Hz is recommended as the maximum bucket size, particularly for the HF bands. (Smaller would be better, but would be too slow.
- 500Hz buckets need 2ms of signal to process. Averaging over 10 samples needs 20ms tones to be transmitted. As this equates to approximately 20cm of movement at a full run, if more error can be tolerated during motion, this is probably the maximum acceptable error. However, a wider FFT bucket, giving faster times, would suffer more from interferers within the chosen bucket in the HF bands.
- The typical TCXO clock drifts over temperature and age for the HF band are given below. When designing the final product these will need to be taken account of when assessing the accuracy of the final system.

Mobile TCXO could be	+/-1ppm ageing in first year
	+/-2ppm temperature –20 to +70°C
	+/-0.5ppm initial set-up
BS OCXO could be	+/-0.05ppm ageing in first year
	+/-0.005ppm temperature –20 to
+70°C	
	+/-0.01ppm initial set-up
Total	+/- 3.565ppm

2.5 Base station identification & calibration

2.5.1 BS identification

Tones from different base stations need to be distinguishable. Although data modulation onto the tones could be considered, it will corrupt an instantaneous phase measurement unless more complicated design counter-measures are taken. A simpler system would use either time or frequency coded identification and a lookup table. These could take the form of a TDMA or FDMA based technique, to have base stations transmitting in specific time slots or on specific frequencies. The choice between the two is application specific as it will be affected by the number of base stations, available bandwidth, duty cycle restrictions and required transmit duty cycles. FDMA will occupy more spectrum slots, albeit within the bandwidth of the maximum tone separation, whereas TDMA will require re-alignment of the individual measurements on each base station before they can be combined into a location algorithm.

2.5.2 System calibration

If the transmitters and receivers in a phase location system aren't synchronised i.e. frequency/phase and time locked, with known start phases for the tones, the calculations fail unless a calibration process exists to provide additional information for the calculations.

Synchronisation could be achieved by a third party system, e.g. GPS, cable etc according to system scale. Any errors in the third part systems will result in a reduced accuracy in the location system. An unsynchronised system could be implemented by placing a receiver at a known location to act as a reference for all other "mobile" receivers. As it receives the same signals as the other mobiles and the system knows its location, information on frequency offset / drift etc can be collected for the base station signals. This can then be broadcast to the mobiles (alternatively via some of the base stations) for inclusion into their location algorithms. This is currently implemented in the Racetrace system.

For a FDMA base station ID structure, the offsets this will cause in the BS-BS tone phase measurements will be accounted for provided the channel separations are small enough for the calibration loop to process the phase twist quick enough.

2.5.3 Interfering signals

In order to work around interfering signals, particularly within the narrow HF bands, a slow time hopping system where stations shift to a different channel for each tone transmission will ensure that all stations get received at least some of the time, providing there are some quiet channels. A 1KHz channel spacing could

be considered as this allows an FFT bin spacing between channels (at HF) to aid channel filtering implementation.

As this system would also guarantee that a station would collide with an interferer if one were in the band, there must be sufficient "spare" base stations being received successfully for the location algorithms to work, i.e. at least 3.

3 PROPAGATION AND PATHLOSS CALCULATION

To provide Racetrace coverage in a typical training village and within buildings it is necessary to determine the line-of-sight propagation range between transmitters and receivers. Sometimes the line-of-sight may be obscured by a building but the received signal level may still be sufficiently high to be useful since the attenuation of signals through masonry blocks is finite. In addition, to provide full coverage in a village a large number of transmitters may be required. However, the maximum number of transmitters that can be located in a single village must be restricted to prevent co-channel and adjacent channel interference.

This section provides estimates of the maximum number of transmitters and line-of-sight and obscured line-of-sight characteristics that can be used in a model to calculate Racetrace coverage in a typical village and hence estimate suitable positions for transmitters.

3.1 Racetrace-Based Capacity Calculations

Suggested frequency bands for licence exempt Racetrace-based location are defined in Section 2 and summarised below in Table 2.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Max. Bandwidth (MHz)	0.03	0.014	0.04	1.75[1]	0.25	0.3	83

Table 2: Possible frequency bands for a Racetrace based system.

In performing the capacity calculation the following assumptions have been made:

- It is not possible to re-use frequencies in different parts of a village because the probability of co-channel interference will be too high.
- 10 consecutive transmissions are required to improve the phase measurement accuracy (see Section 2)
- The maximum error when moving at 10m/s must be no greater than 20cms. This equates to a maximum measurement time of 20ms. With 10 chirps for each measurement no chirp must be longer than 2ms.This equates to a minimum bucket size (channel bandwidth) of 500Hz.
- The update rate must be at least 5 times per second
- The capacity calculations assume a worst case scenario where there is no obscuration by buildings and the receiver is approximately equidistant from each of the transmitters.

[1] The 433MHz band has an available bandwidth of 1.75MHz. However, the use is restricted in some of the band by duty cycle and power. 434.04-434.79MHz has 10mW max. EIRP, 100% duty cycle and 25kHz maximum channel bandwidth.

The two methods considered here for separating transmissions are FDMA (Frequency Division Multiple Access) and TDMA (Time Division Multiple Access).

3.1.1 FDMA

FDMA can be used when there is sufficient bandwidth to separate the buckets/channels. Table 3 provides an estimate of the bucket size and number of buckets possible in each frequency band. The minimum bucket size is limited by the need to minimise chirp length when the receiver is moving as described above. The minimum bucket size in some bands is also limited by the maximum number of buckets that can be processed at the receiver. The total number of buckets in a band must not be so high as to make the FFT implementation impractical. A maximum FFT size of 1024 points is assumed.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7 -870	2400
Max. Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Bucket size (Hz)	500	500	500	1462	500	584	161,793
FFT size (pt)	128	64	256	1024	1024	1024	1024
Chirp time (ms)	2	2	2	0.7	2	1.7	0.006
Chirp time x10 (ms)	20	20	20	7	20	17	0.06
Error when rx is moving (cm)	20	20	20	7	20	17	0.06

Table 3: Bucket size and no. of buckets

Theoretically, there are a large number of buckets available in each band. However, there are number of practical reasons why the number of channels that can be used could be significantly less than this:

- The FFT has finite selectivity. An FFT has approximately 13dB adjacent channel selectivity without windowing. A windowed FFT will be much higher and will enable the adjacent channel +1 to be used.
- The dynamic range of the A-D may limit the amount of adjacent channel rejection possible. However, fast 14 or 16 bit A-Ds are now available. Therefore, this is a performance vs. cost consideration.
- Multipath can be mitigated by transmitting a signal spread signal over the available band instead of transmitting a single tone.

However, if FDMA is possible then a Racetrace tracking system may be implemented without using TDMA.

Table 4 provides the maximum capacity assuming that FDMA is possible and tones (rather than spread or modulated signals) are transmitted. The number of useable buckets/channels is derived assuming that adjacent channels are not used. This ensures that the receiver filtering is practical. If each transmitter operates on a different

frequency the maximum number of transmitters is equal to the number of useable channels. In the frequency bands where the capacity is very high it may be more appropriate to widen the buckets to reduce the error due to motion. The bucket size may also be increased to simplify the receiver filtering.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65[2]	869.7-870	2400
Max. Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Bucket size (Hz)	500	500	500	1462	500	584	161,793
No. of buckets/ channels	60	28	80	513	500	513	513
No. of useable channels	30	14	40	257	250	257	257

Table 4: Maximum capacity assuming FDMA is used (but no TDMA)

FDMA has one specific advantage over TDMA. This is that all FDMA transmissions occur simultaneously. Using TDMA it is necessary to include an additional processing function to refer all TDMA transmissions to the same point in time.

3.1.2 TDMA

Transmitters can be either unsynchronised or synchronised by a timing reference such as that available from GPS (assumes GPS coverage to the base stations is available). The capacity for both scenarios will be calculated.

The capacity of a TDMA system is also related to the chirp length and hence the bucket size. There are two extremes for the bucket size. The minimum bucket size for a practical implementation is shown in Table 4. The maximum bucket size is the equal to the bandwidth of each allocated band but will make the receiver susceptible to in-band interference from other sources.

For a synchronised system the maximum number of transmitters when using TDMA only is simply related to the chirp length and update rate. Table 5 shows the maximum number of synchronised transmissions without collisions when using the minimum bucket size. There is a direct relationship between bucket size and the capacity. For example, if the bucket size is increased by a factor of 10 then the capacity will increase by a factor of 10.

[2] Duty cycle of transmitters must be <10%

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Update rate (per sec)	5	5	5	5	5	5	5
Chirp time x10 (ms)	20	20	20	7	20	17	0.06
Max. no. of txs	10	10	10	28	10	11	3333

Table 5: Maximum number of transmitters using TDMA in a synchronised system (minimum bucket size)

For an unsynchronised system there will always be a finite chance of a collision. A CSMA algorithm could be used but at the expense of extra complexity at each transmitter in the form of an additional receiver.

The probability of a collision is related to the required update rate and chirp length. For example, in the 6.7MHz band the 200ms period between updates can be divided into ten 20ms timeslots. Since any overlap between two transmissions is a collision the probability of collision for two transmitters is 0.3. The probability of collision can be calculated for any number of transmitters. Table 6 shows the probability of collision for N=2 to 5 transmitters in each of the 7 bands proposed.

The probability of collision is reduced when the bucket size is increased and the chirp time reduced. Table 7 shows the effect on the probability of collision by increasing the bucket size and reducing the chirp time by a factor of 10, for example.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Update rate (per sec)	5	5	5	5	5	5	5
Chirp time x10 (ms)	20	20	20	7	20	17	0.06
Prob. Of collision (2 TXs)	0.3	0.3	0.3	0.105	0.3	0.255	0.0009
Prob. Of collision (3 TXs)	0.66	0.66	0.66	0.28	0.66	0.59	0.003
Prob. Of collision (4 TXs)	0.88	0.88	0.88	0.49	0.88	0.83	0.005
Prob. Of collision (5 TXs)	0.97	0.97	0.97	0.67	0.97	0.85	0.009

Table 6: Maximum no. of transmitters using TDMA in an unsynchronised system and probability of collision (minimum bucket size)

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Update rate (per sec)	5	5	5	5	5	5	5
Chirp time x10 (ms)	2	2	2	0.7	2	1.7	0.006
Prob. Of collision (2 TXs)	0.03	0.03	0.03	0.0105	0.03	0.0255	0.00009
Prob. Of collision (3 TXs)	0.09	0.09	0.09	0.03	0.09	0.07	0.0003
Prob. Of collision (4 TXs)	0.17	0.17	0.17	0.06	0.17	0.14	0.0005
Prob. Of collision (5 TXs)	0.26	0.26	0.26	0.1	0.26	0.23	0.0009

Table 7: Maximum number of transmitters using TDMA in an unsynchronised system (x10 minimum bucket size)

3.1.3 FDMA and TDMA

Clearly, a combination of synchronised TDMA and FDMA can be used to yield maximum capacity. This is shown in Table 8 for the minimum bucket size. Clearly, if the bucket size is larger the proportion of TDMA and FDMA channels changes but the total capacity remains the same.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65[3]	869.7-870	2400
Max. Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Bucket size (Hz)	500	500	500	1462	500	584	161,793
FDMA capacity	30	14	40	257	250	257	257
TDMA capacity	10	10	10	28	10	11	3333
Combined capacity	300	140	400	7196	2500	2827	856581

Table 8: Maximum capacity using synchronised TDMA and FDMA.

[3] Duty cycle of transmitters must be <10%

3.2 RACETRACE-BASED PROPAGATION CALCULATIONS

Estimates have been made of line-of-sight propagation for a village excluding the obscuration effects of the buildings for each of the frequency bands. These are shown in Table 9.

Parameters used in the calculation for each band include

- Maximum allowed transmit power
- Antenna efficiency
- Receiver sensitivity (derived from bucket bandwidth)
- Assumes a signal to noise ratio requirement of 20dB for 5.7° phase accuracy rms
- Assumes receiver NF (noise figure) = 9dB
- Assumes antenna is placed on top of head so that there is no absorption by the body or head.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Bucket size (Hz)	500	500	500	1462	500	584	161,793
Transmit power (dBm)	3 (≡42dBuA/m in far field)	-4 (≡42dBuA/m in far field)	10	10	27	7	10
Receiver sensitivity (dBm)[6]	-117	-117	-117	-112	-117	-116	-92
Transmit antenna gain (dB)	X[4]	X[4]	0[5]	0[5]	0[5]	0[5]	0[5]
Receive antenna gain (dB)[6]	X[4]	X[4]	-20	-10	-7	-7	-3
Max. tolerable loss (dB)	X[4]	X[4]	107	112	137	116	99
Max. LOS distance – theoretical (km)	X[4]	X[4]	131	22	195	17.3	0.9

Table 9: Maximum LOS range / band

To build a propagation model of the village to assess coverage it is necessary to know the loss through building material (masonry block) and calculate the maximum tolerable loss before the signal to noise ratio at the receiver is less than 20dB. Measurement of loss through building materials have been made by the U.S.National Institute of Standards and Technology (NIST). 'Construction Automation Program Report No. 3' provides details of Electromagnetic Signal Attenuation in Construction Materials. The loss through 203mm thick masonry block is shown in Table 10.

[4] The distance calculations for 6.7MHz and 13.5MHz have not been performed because of the difficulty of assessing signal strengths in the near field. The calculation is further complicated by the fact that the receiver antennas will be very small compared to wavelength. The radiation pattern especially in the near field will be very complex. It is suggested that the propagation characteristics are derived using data for 6.7 and 13.5MHz should be derived empirically using representative antennas.

[5] Cannot be greater without exceeding EIRP limits

[6] Assumes 20dB required S/N and 9dB receiver noise figure in the calculations below.

Receiver sensitivity calculation:

Receiver sensitivity = 10^4log(kT) + rx noise figure + required S/N + 10log(bucket size)

Receiver antenna gain:

It is assumed that the receiver antennas need to be unobtrusive when carried and therefore small. In most cases the antenna size will be too small to be ideal and may been quite inefficient. The antenna gain has been estimated by calculating the maximum gain achievable for an antenna with one of its dimensions as 10cms. Since this is an ideal antenna a further 10dB is subtracted from the calculated gain figure to account for inefficiency and non-ideal implementation.

Frequency (MHz)	6.7	13.5	40.6	434.04-434.79	869.4-869.65	869.7-870	2400
Bandwidth (MHz)	0.03	0.014	0.04	0.75	0.25	0.3	83
Loss in masonry block (203mm thick)	X[7]	X[7]	X[7]	7	11	11	13

Table 10: Attenuation through masonry block

These figures can be used in the following equations to calculate the loss along the propagation path.

Free space loss (FSL) = 20log f (in MHz) + 20log d(in cms) – 67.5

To calculate the signal strength at any distance from a transmitter:

RSL = Tx power (dBm) + Tx antenna gain (dB) + Rx antenna gain (dB) + [20log f (in MHz) + 20log d(in cms) – 67.5] – att/block * no. of masonry blocks traversed

Whenever, the RSL falls below the sensitivity given in Table 9 then the S/N ratio at the receiver will be <20dB. Antenna gains are also provided in Table 9.

To provide a guide to either the possible range through a given number of walls or the number of walls penetrable for a given range, figures from Table 9 & Table 10 are combined to produce Table 11. This includes floor attenuation figures extrapolated from several published sets of measurement results.

Frequency (MHz)	434.04-434.79	869.4-869.65	869.7-870	2400
Max. tolerable loss (dB)	112	137	116	99
Max. LOS distance – theoretical (km)	22	195	17.3	0.9
Range including 1wall	9782	54814	4880	198
Range including 2walls	4369	15449	1375	44
Typical floor loss	25	30	30	35
Range for 1 floor	1231	6150	548	16

Table 11: Range through walls and floors

[7] There is no commonly available data for attenuation through construction materials at 6.7, 13, and 40.6MHz. It is suggested that the propagation characteristics are derived using data for 6.7, 13.5 and 40.6MHz should be derived empirically using representative antennas in a typical environment.

As can be seen, for the 2.4GHz band, base stations should be able to penetrate the external wall of buildings up to approximately 200m away. This allows cover for the village diameter. More realistically, nearer base stations with be used, and within ~40m of the mobile in a building, they should be able to penetrate both the external and 1 internal wall. This is also shown in Section 4. Note that floor penetration will be much more difficult.

4 URBAN RF COVERAGE

To investigate a typical deployment within a military combat training area it is necessary to build a model of a village and deploy a simulated radio transmitter network. The coverage from direct line of sight paths can then be calculated by using the propagation characteristics given in Section 3.

4.1 Model description

Figure 4 shows a typical village where multiple houses are distributed around two main streets. The village is 190m by 160m which represents a large military training area of this type. The houses were assumed to be made of breeze blocks and of varying shape and height. The average house is about 10m by 15m and 6m high (8m at the apex of the roof). It was decided to model the transmitters at the three highest frequencies (see Section 6).

Modelling of the transmitters was done, assuming omni-directional antennas with appropriate opening angles, at each site. Masts of different height were deployed by attaching them to the side of some of the houses in the village. The red crosses in Figure 4 represent the transmitters that are 15m above ground level and the blue crosses represent the transmitters that are 22m above ground level (usually on the roofs of tall buildings). After modelling it was decided that 20 transmitters were required to give sufficient coverage across the whole village at each frequency. Each of these 20 transmitters is shown in this Figure.

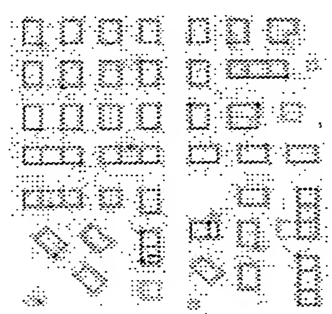


Figure 4: Plan of a typical village for a military combat training area. These consist of houses of differing heights, sizes and internal structures. Also shown is a typical deployment of 20 radio transmitters.

The propagation from each transmitter was modelled by looking at the line of sight from the transmitter to the location requiring a fix. The power loss (in dB) was calculated along the line of sight, taking into account the propagation through air and through the walls. It was assumed that the transmission power at each frequency was 10dB. At each location the final dB level for the transmitter was then checked to see if

it was above the threshold for detection (see Section 3) and the number of transmitters seen at that location incremented if it was.

4.2 Results

Figure 5 shows the results of the propagation calculation for the village shown in Figure 4. The frequencies shown are 433MHz, 868MHz and 2.4GHz. To obtain a location the transmitter must be able to see 3 or more transmitters through a direct line of sight. The direct line of sight includes through air and wall propagation statistics. This is shown as the blue region in the map. The green region in the map can detect line of sight signals from, on average, 10 transmitters. This results in a significant increase in the accuracy and gives a good chance of being able to eliminate multipath effects. The red region in the map can detect almost all of the transmitters in the village through line of sight deployment. The black regions show where no location is possible (less than three light of sight signals).

As can be seen from this figure 20 transmitters is enough to cover the village for a system operating at 2.4GHz. At 433MHz is seems that 20 transmitters is too many. Almost every transmitter can be seen at almost every location around the village as the 433MHz signal is much better at propagating through concrete walls than the higher frequencies. Figure 6shows the coverage for a deployment of 8 transmitters around the village. As can be seen from this plot 8 transmitters is adequate to give coverage at 433MHz whereas at 2.4GHz a lot of the village is not covered. Therefore, between 8 and 20 transmitters are needed to give coverage for a typical village dependent on the frequency used.

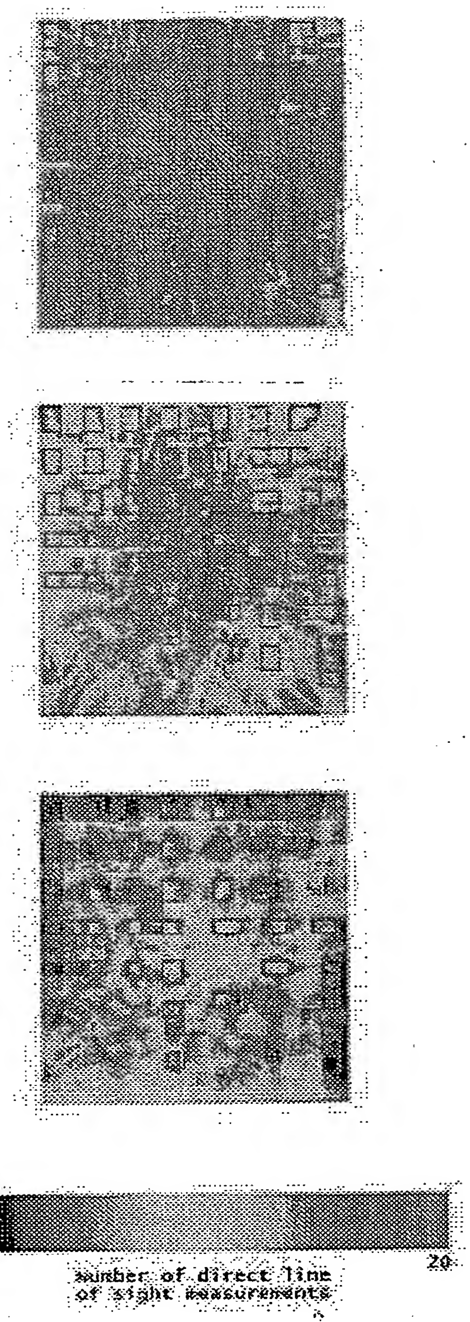
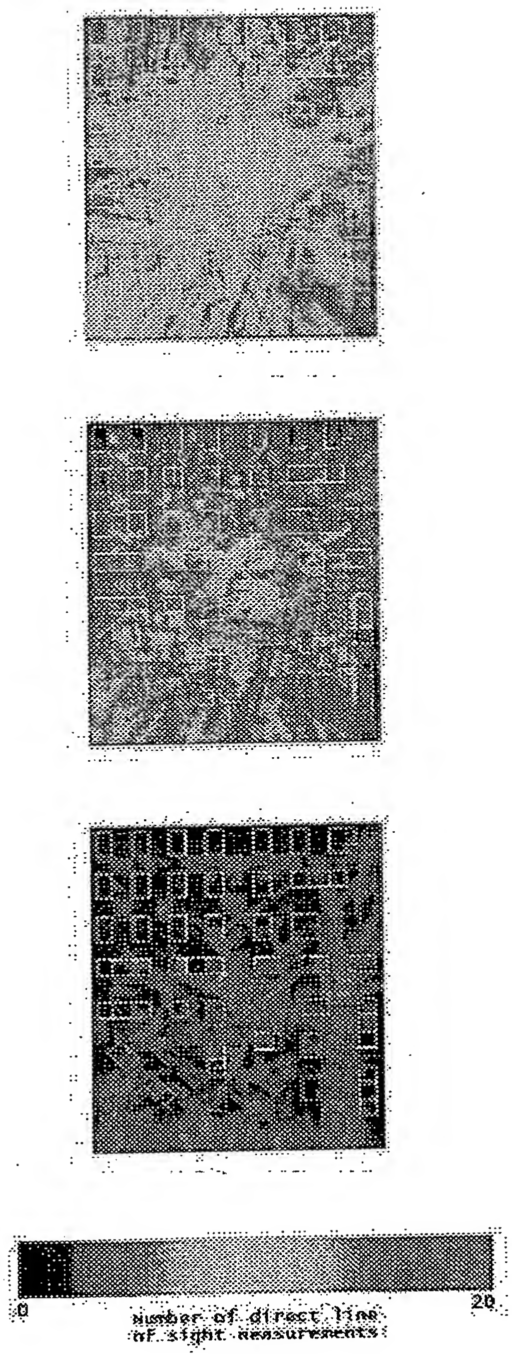


Figure 5: Coverage plot for the village shown in Figure 4 at each of the three highest frequencies. A total of 20 transmitters were used in each of the above.



433MHz

868MHz

2400MHz

Figure 6: Coverage plot for the village shown in Figure 4 at each of the three highest frequencies. A total of 8 transmitters were used in each of the above.

5 THE MULTIPATH ENVIRONMENT

This document summarises various multi-path environments and mitigation techniques for use in the licence exempt bands. The environment is indoor and urban-outdoors, in small village sized sites. The current application is for a personnel tracking system using phase measurement based systems.

5.1 General multipath

This section describes what multipath is and provides definitions for the common terms used to quantify it.

The term "multipath" covers the effect the physical environment has on radio waves when there are reflective surfaces present for the transmitted waves to bounce off and back to the receiver, as in Figure 7.

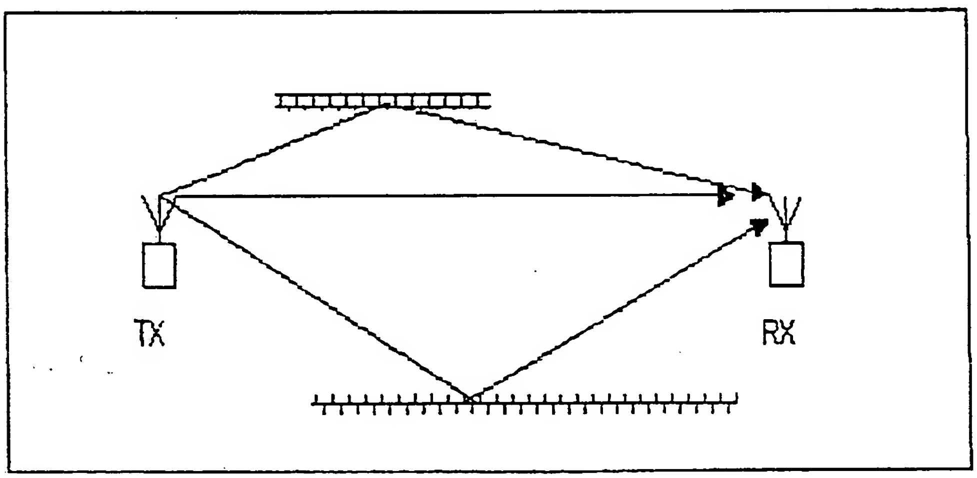


Figure 7: Example multipath components.

This results in signals arriving with different delays due to the different distances travelled, corresponding to phase differences between all the arriving paths. All the signals add together either constructively, destructively or somewhere in between with the effect of distorting the original signal. For small reflected signal powers, the distortion takes the form of adding a phase noise component. For reflected signals that approach the power of the LOS path, the destructive interference produces deep fades in signal strength.

With a LOS path, these fades are referred to as Rician fading due to the probability of the fades following a Rician characteristic. If there is no LOS path, this leads to a Rayleigh probability characteristic and Rayleigh fading. See Figure 8. Although vital for communications systems, for the purposes of phase based location systems, these amplitude based characteristics are less important and so are not covered further

here. Of more concern is the phase error that multipath components introduce. Hence we will concentrate more on characteristics in the time domain than amplitude domain.

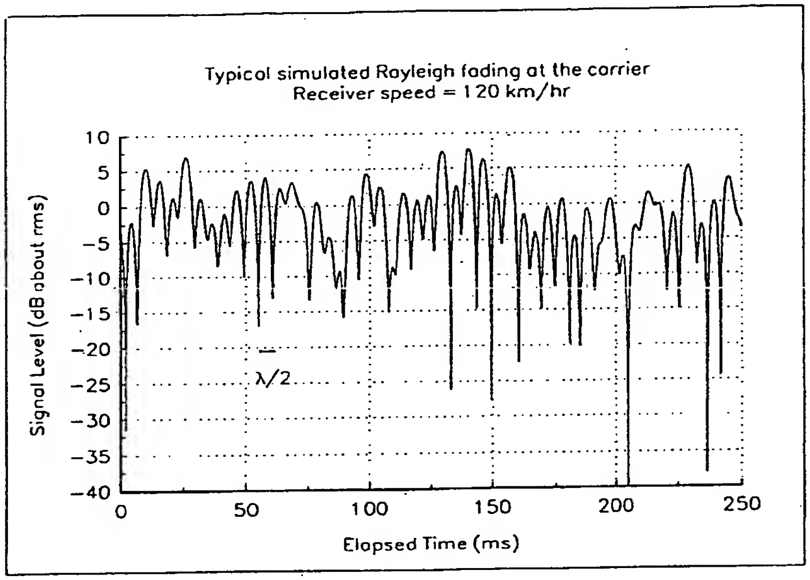


Figure 8: Typical Rayleigh fading channel at 900MHz

5.1.1 Multipath time dispersion parameters

Several key parameters are used to characterise a multipath channel in the time domain. These originate from looking at the time domain impulse response of the channel, with all measurements made relative to the first detectable signal (the direct path in Rician channels). As the timescale measurement resolution cannot be made infinitely small, it is usually broken into resolvable timeslots whereby any paths arriving within the same timeslot cannot be separated and averaged, but those that are outside the timeslot can be resolved as separate paths.

Averaging individual instantaneous power delay profile measurements over the local area generates a power delay profile of the channel. Parameters taken from this profile are then used to describe the channel. An example is shown in Figure 9.

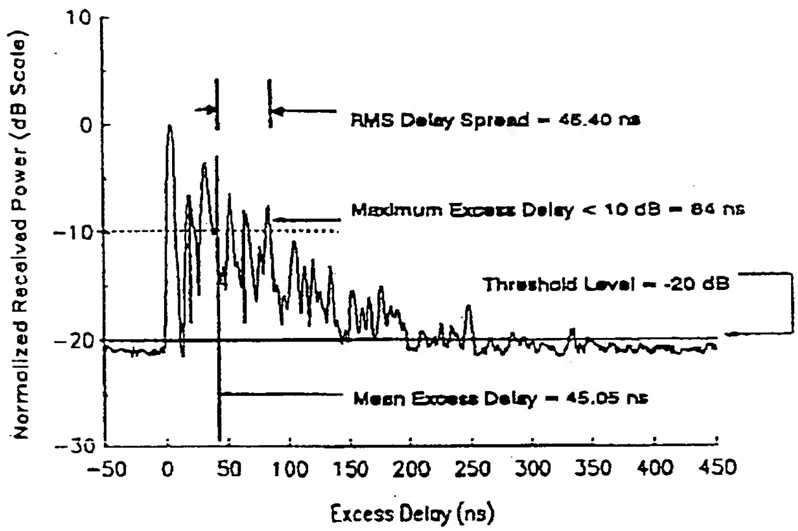


Figure 9: Example indoor power delay profile

Excess Delay τ_i

The delay between the first arriving component and the specific i'th component.

Threshold level XdB

The level relative to the maximum signal (not necessarily the first), below which any signals are ignored.

Maximum excess Delay τ_X

The maximum excess delay producing a signal above the Threshold level. Sometimes refered to as **excess delay spread**. XdB should be stated.

Mean excess delay $\bar{\tau}$

This is the first moment of the power delay profile. ie.

$$\bar{\tau} = \frac{\sum_k P(\tau_K)^* \tau_K}{\sum_k P(\tau_K)}$$

component power levels $P(\tau_K) = a_K^2$

RMS delay spread σ_τ

This is the square root of the second moment of the power delay profile ie.

$$\sigma_\tau = \sqrt{[\overline{(\tau^2)} - \bar{\tau}^2]}$$

where

$$\overline{(\tau^2)} = \frac{\sum_k P(\tau_K)^* \tau_K^2}{\sum_k P(\tau_K)}$$

5.2 Typical Radio Environments

5.2.1 Indoors

Initial measured values for delay spread have been taken from Rec.ITU-R P.1238.1 and are shown in Table 12.

Frequency /MHz	Environment	RMS delay spread σ_τ /ns		
		Low. Not extreme	Median. Frequent	High. Extremely rare
1900	Residential	20	70	150
1900	Commercial	55	150	500
1900	Office	35	100	460
5200	Office	45	75	150

Table 12: Typical measured values of Indoor RMS delay spread

As base stations and mobiles may be in closer proximity to each other and the walls than would normally be the case for an "office", it is suggested that as little as 2ns excess delay could be experienced.

At the other extreme, the maximum excess delay will be dependant on the through wall attenuation, reflection coefficient for reflected signals and the room size. For high attenuation materials such as concrete, reflections from adjacent rooms are unlikely to affect measurements, as they will be at too low a level. This will limit reflections to within the room. Given the standard size room in houses (approximately 5m x 5m), and the limited number of reflections that can occur before the multipath signal level is reduced to insignificance, it is proposed that 160ns be absolute maximum excess delay to be considered relevant.

An in-room propagation model has been constructed to show the delay profile between a transmitter and receiver within the room. A sample output in the 2.4GHz band, for a 10m x 5m concrete room with the Tx, Rx coordinates of (5m, 2.5m) & (9.9m, 4.9m) is shown in Figure 10. The short path reflections due to the receiver proximity to the corner and the reflection spread matching the room aspect are clearly shown.
The results were also integrated over all possible positions within the room to indicate the probability characteristics of the delay spread and later, used to model phase measurement based location errors. This is all described in detail in Section 8.

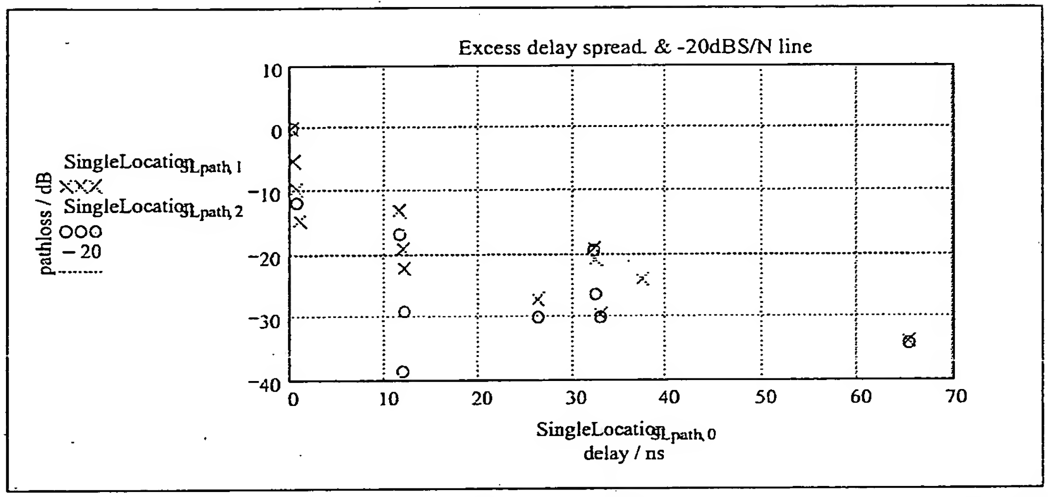


Figure 10: Multi-path delay spread for 1 location in a room. "x" & "o" indicate vertical & horizontal polarised power with respect to LOS arriving at receiver

5.2.2 Outdoors

Initial measured values for delay spread have been taken from Rappaport[8] and are shown in Table 13.

Environment	Frequency /MHz	RMS delay spread σ_{τ} /ns	Notes
Urban	910	1300 avg, 600 st dev, 3500 max	New York city
Urban	892	10-25us	Worst case San Francisco
Suburban	910	200-310	Average typical case
Suburban	910	1960-2110	Average extreme case

Table 13: Typical measured values of Outdoor RMS delay spread

Although the average excess delay will be larger than indoors, as the personnel could still be very close to walls, the excess delay for the first bounce path could still be as little as a few ns. However, 10s to 100s of ns are more likely for most of the paths. Also, the accuracy requirement for positioning is unlikely to be as tight as that for indoors, so the importance of the close in delays is similarly reduced.

As there are no intervening walls to add attenuation, the maximum excess delay will depend on the ratio of LOS to reflected pathlength, the S/N requirement and the limit of sensitivity of the receiver. The worst scenario is a transmitter at the limit of range for

[8] Rappaport T.S. Wireless Communications: Principles and Practice 1996

the receiver and a reflector providing a multipath component just above the S/N threshold. This would typically come from reflectors well outside the zone of operation and hence can be difficult to control. It is proposed that 2-3000ns be treated at the range of excess delays with the average being around 50-300ns.

Based on the same model structure, treating the road as a long thin "room" for the purposes of 2D reflections, Figure 11 shows the results of a specific delay spread at 2.4GHz of a 15m wide by 200m long with Tx, Rx coordinates of (150m, 14.9m) and (50m, 10m). As is seen, even for the size of the area, very short multipath is still present at this location. Mitigation techniques would need to consider this.

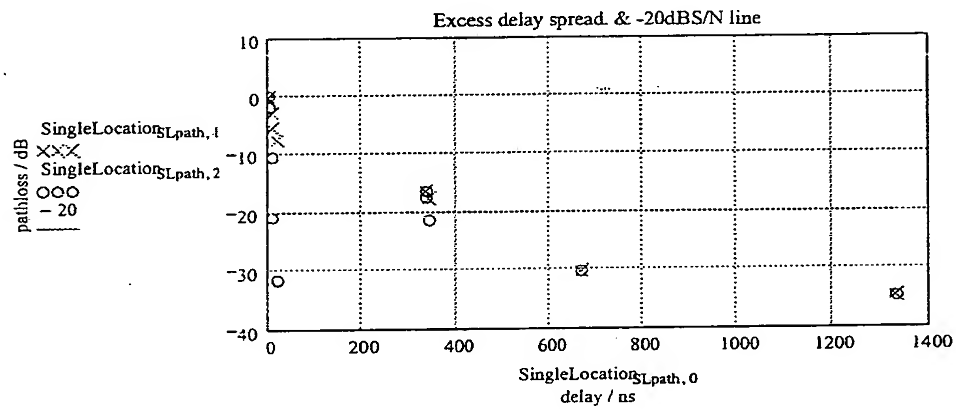


Figure 11: Multi-path delay spread for 1 location in a street. "x" & "o" indicate vertical & horizontal polarised power with respect to LOS arriving at receiver

As for the indoor scenario, this information was integrated over all possible positions on the street to indicate the probability characteristics of the delay spread and later, used to model phase measurement based location errors. Again, this is described in detail in Section 8.

5.2.3 Outdoor to indoor

In order to assess coverage indoors from external base stations, the building penetration needs to be calculated. Whilst detailed prediction also needs the angle of incidence etc just as for the indoor propagation model, a first order model can use typical attenuation figures for the materials to calculate the signal level entering the room and then use an indoor multi-path model after that. This subject is discussed in Section 3 and Section 4.

6 MULTIPATH MITIGATION TECHNIQUES

So far only phase measurements on carrier tones have been discussed. These will obviously be affected to varying degrees by multipath at the different frequencies. Separate investigations covered how to mitigate multipath to a sufficient degree for the described core location system to work. The techniques investigated are covered below.

6.1 Multipath Mitigation techniques

6.1.1 Antennas.

There are number of things that can be doe with modified antenna systems to combat the multipath problem.

Diversity / redundancy

It is quite common in fixed and mobile communications links to employ receiver diversity. In its simplest form you combine the outputs of multiple receive antennae to one receiver. The antennae are positioned so that when one of them is in a deep fade the other is not. This ensures the combined output, whilst at a lower overall level than that of a single antenna, rarely suffers from deep multipath induced fades.

For a location system this approach has the problem that we are deliberately introducing an uncertainty in the position of the receive antenna. (We never know which antenna the dominant signal was received through.) This can be overcome by

- Ensuring that the uncertainty is insignificant by making the antennas spacing significantly less than the required location accuracy. Note that this is inconsistent with a carrier phase measuring location system where the separation would have to be $>\lambda/2$ for diversity and $<<\lambda$ for location accuracy!
- Ensuring that only a single receive antenna used for each measurement and that any offset between antennae is corrected for in the location calculation.
- Processing the signal from each antenna separately with a complete receiver chain for each antenna. This allows very large spacings to be used but does increase cost because of the duplicated receiver chain.

With all diversity systems there is the problem of deciding whether a received measurement carries good or bad (direct or reflected path) information. In theory diversity will only help a location system where one of the receive antenna is in an obstructed path rather than a faded one. Since if we have a fading problem we have multipath signals and we are unable to distinguish the direct from the reflected paths.

Polaristion

Polarisation is a very useful tool for rejecting reflected signals because on reflection the radio wave is reversed (just like the reflection of an image in a mirror). This allows us to distinguish direct from single reflected signals in some cases by observing their

polarisation. For horizontal and vertical polarised waves the reversed polarisation (if from a vertical reflector) looks the same as the original polarisation to a receive antenna. Circular polarisation is required to ensure that original and reflected waves always look different to a receiver.

For location systems we often require omni-directional receive antennae (because by definition we don't know a priori where the signal is coming from). Circularly polarised omni-directional antennae are difficult to manufacture and therefore much larger and more expensive than linearly polarised ones. Nevertheless polarisation is an effective tool to deal with reflected multipath.

Note multipath problems can also be introduced by refraction of the waves and polarisation is not effective in these cases. Refraction normally only occurs on long propagation paths. (e.g. on satellite links).

Directionality

As noted above directional antennae are generally a problem for use on the target (object to be located) in a location system because of the lack of knowledge about where the object to be located is.

Sectored receive antennae can be used with diversity receivers to provide some protection against multipath. This works if the direct and reflected rays arrive from different directions and so the destructive interference that would occur at an omni-directional antenna can be avoided. These should be particularly useful in indoor environments where with the short propagation distances the reflected path is more likely to arrive from a different direction. There is still the problem of identifying which signal was the direct and which was the reflected path.

They can be used on fixed stations and particularly on the edges of the location area where they will remove or reduce the effect of reflectors outside the location area.

6.1.2 Smart antennas.

Beamsteering

Beamsteering antennae overcome the main disadvantage of using directional antennae. Whilst simple beamsteering systems work at RF and need a priori knowledge of the direction required modern systems allow a single array antenna to use multiple beams simultaneously and in the case of a receiver to form the beams on the basis of the received signal. A smart steerable antenna could provide multiple steerable narrow beams that track all transmitters simultaneously and provide rejection of reflected signals. These systems are however still very expensive to construct.

This approach would require a multi element array antenna with consequent increase in antenna size.

Null steering

An interference nulling smart antenna which is a variant of beam steering could reject an interfering signal. This relies on the null that formed with beam-steering algorithms

can theoretically be infinitely deep whereas the maximum gain over a single element antenna is only the number of elements used (e.g. 6dB for 2 elements).

This approach allows a 2 element antenna to remove one multipath signal without significantly changing the wanted signal gain. Again the problem is identifying which is the unwanted signal in the first place rather than the signal processing algorithms.

A variation on this technique is to transmit a null in on direction from the transmitter and then change that direction with time creating a rotating null in the field around the transmitter. This allows a receiver to examine the envelope of the signal received and identify whether it is subject to multipath effects from the number of dips in signal level that are observed in the period of one rotation. This would then provide a quality factor for any range deduced from that signal. Indeed if the rotation wan synchronised it would allow a range and bearing from the transmitter to be deduced by the receiver.

Adaptive polarisation

Another smart antenna system could be designed to use cross polar antennae. Then by using the adaptive filter it could track the polarisation of the dominant received path and ignore cross polar signals. This approach only works with polarisations that can be distinguished from their reflected selves. Linear polarisations can look the same when reflected (depending on the angle of incidence with the reflector), circular polarisations always looks different.

Note that the overall effect of this with linear polarisations is similar and inferior to the use of circular polarisation. Its performance is worse than using circular polarisation because of the dependence on the angle of the reflector but the antenna can be physically smaller than a circularly polarised one.

All the smart antenna solutions would need a multichannel receiver with matched delay and phase characteristics. This in itself is a nontrivial problem.

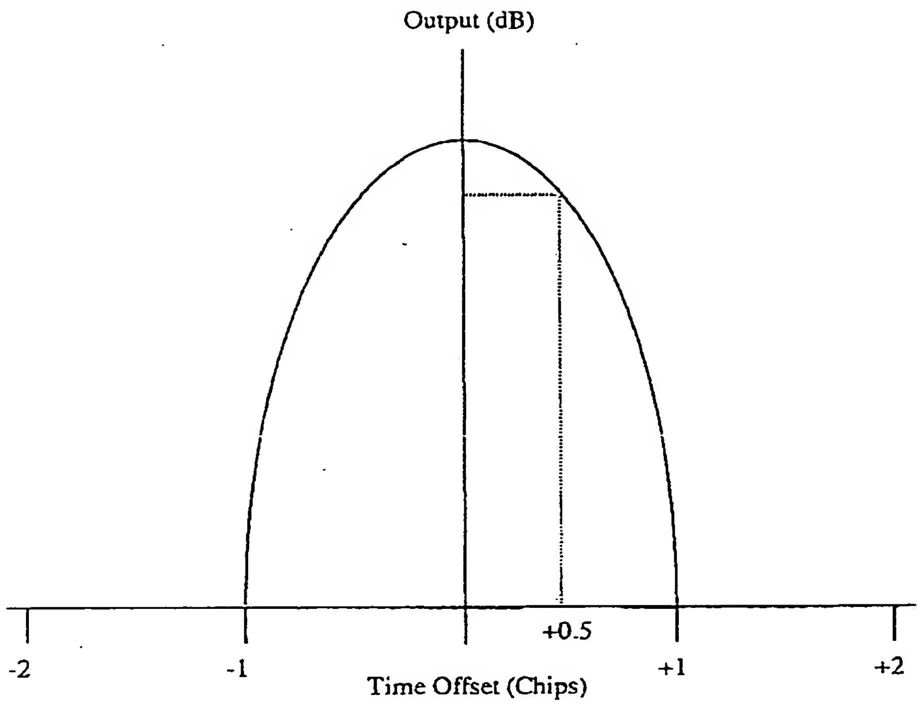
6.1.3 Frequency spreading.

Direct Sequence spread spectrum (& Frequency sweep)

Direct Sequence Spread spectrum is used for data communications in unlicensed bands. It is attractive because it allows different users to be separated by their transmission time rather than frequency. The use of Direct Sequence Spread Spectrum could in principle allow us to separate and reject all multipath components. However receiver practicalities get in the way.

Suppose we need a positional accuracy of 0.5m. We therefore need to reject a multipath signal with more than 0.5m path difference from the direct signal. This implies rejecting a signal arriving 1.7ns later. To reject something in the DSSS despreader we need it to have a time offset of greater than 1 chip period (See Figure 12). This implies a chip rate of 600MHz for our 0.5m system.

This system will require greater bandwidth than is available in any of the non specific unlicensed bands below 123GHz.



DSSS Correlator output Vs time offset

Figure 12: Typical DSSS Correlator output

In practise a multipath element of 0.5 chip period will be attenuated by 6dB. At this attenuation a signal delayed by 5ns will only introduce an error of approximately 0.5m which would be on the limit of desired performance. This corresponds to a chip rate of 200MHz which could be accommodated in the 24GHz radio band.

With a chip rate of 41.75MHz (null to null bandwidth filling the 2.4GHz ISM band) we could therefore achieve an accuracy of about 3.6m best case in the presence of multipath.

Frequency hopping spread spectrum

Frequency hopping uses a different approach. For communications this works by assuming that at some frequencies the signal will be lost due to the multipath fading whilst at others the signal will get through. The data has FEC added and is interleaved across the data bursts with the overall effect that the data lost from one burst not being received can be recovered from the good bursts.

For a location system the difficulty is classifying those bursts with good information (direct path) and those with bad information. There is also the issue that frequency selective fading that FHSS overcomes normally will only occur where there is no dominant direct path. If this is this case we cannot use the received information for location purposes as the range it represents will be incorrect. See Figure 13.

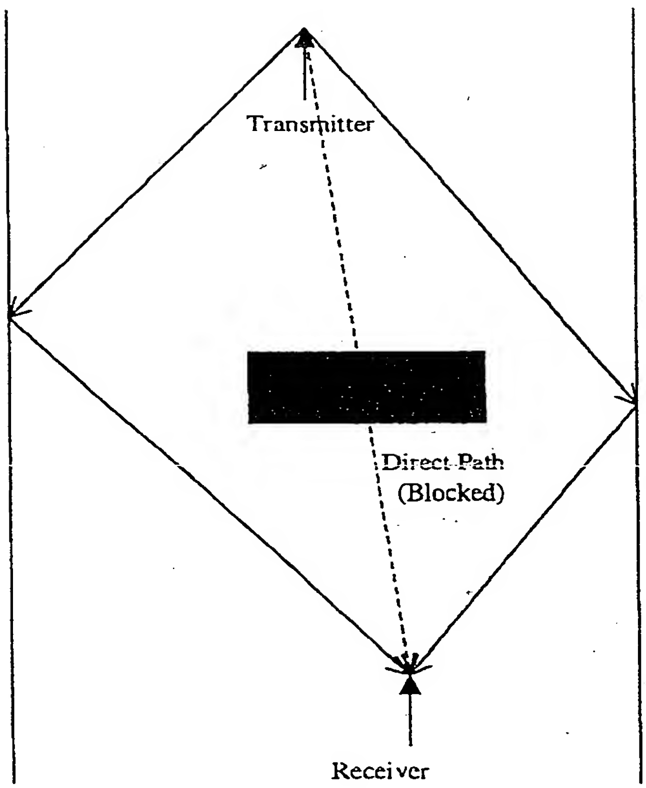


Figure 13: Example of no LOS path presence

It in interesting to note that fundamentally we take a number of measurements that include clean and corrupted measurements. We then need to classify the measurements into good and bad ones, discard the bad ones and yet have enough good ones to achieve the required performance.

6.1.4 Channel equalisation

Channel equalisation when applied to communication channels tries to remove the effects of multipath propagation from the received signal. There are many different techniques but they all essentially perform an inverse filter function on the received signal to recover the transmitted one. This has the advantage that the multipath parts of the signal are actually used to reconstruct the transmitted signal and hence these system often work better than a single line of sight link would. The main problem is that for communications the absolute propagation time is not important so these techniques tend to loose the information. This is a problem for location systems.

Some of the techniques actually measure the impulse response of the radio channel and this will (with sufficient processing bandwidth) allow different multipath delays to be resolved. The main issue here is that the bandwidth required to resolve different paths is very large, exactly the same as for the DSSS resolution. However using this technique we have the opportunity to trade Signal to noise ratio for bandwidth. So for example if we need to resolve 0.5m rather than using 600MHz bandwidth we could

used (say) 60 MHz and then use signal processing to deconvolve the different path responses. The additional signal to noise required is because the deconvolution process emphasises the noise and so we must have adequate SNR to allow the process to work.

Some further work is required to confirm the SNR improvement required but Shannon's law would imply 10dB SNR improvement for a factor of 10 in bandwidth.

It should be noted that the signal processing techniques required to resolve the multipath elements here are exactly the same as required for the base calling work we have been doing within the life sciences area with great success (this is a patented technique).

6.1.5 Reject poor signals statistically

Where a signal is faded we cannot get any information that is useful from it. If it is a frequency selective fade we can change frequency but there is still no useful information because if it is frequency selective we are in a multipath situation.

This demonstrates that poor signal strength is an indicator of bad information but good signal strength is not an indicator of good information. In a situation where we have continuous reception on a narrowband link, monitoring the signal strength and noting the presence of deep fades is a strong indicator of multipath. It is however impossible to be sure that you are receiving a direct path without taking into account other information.

To be confident you are receiving direct path checking its location statistics against other independent location variables is needed. These variables could be obtained from other radio paths and then correlated against one another. In addition correlation against the same path over time and checking for unreasonably fast movement etc.

6.2 Summary

We can classify the above into three categories of technique.

6.2.1 Physical path rejection

Reject the unwanted reflections using physical characteristics that are changed by the reflection process itself. Directional receivers and polarisation separation are strong techniques to use here.

These techniques are effective for all path delays but do have some limitations.

- Directional antennae cannot distinguish small differences in incident direction. The makes them more useful for short direct path length situations (e.g. indoors) where the geometry increases the likelihood of large angular differences between interfering rays.

- Polarisation cannot distinguish signals after even numbers of reflections. This makes it attractive where path length differences are very short and it may need to be used in conjunction with other techniques.

6.2.2 Path rejection in receiver

Separate the different multipath components in the receiver, to do this needs a wideband system (so that paths with small timing differences can be resolved) and signal processing in the receiver to discriminate between the different paths.

This approach is very effective for long path delays where there is a direct path but still has limitations:

- It can only identify the direct path by the first to arrive at the receiver. If there is no significant direct path it cannot be identified and the first indirect path will be selected.

6.2.3 Statistical analysis of redundant paths

It is possible to make a large numbers of measurements on many paths and identify in the receiver whether a signal is likely have taken a direct or indirect path and ignore those unlikely to be direct ones. This approach needs a large number of extra transmitter/receiver stations so that there will be at least 3 good independent variables to allow a triangulation to be performed.

Trials on this sort of technique in the mid 1990s around Cambridge suggested 7 paths were needed for reliable performance on long MF transmission paths.

6.3 Recommendations

To combat multipath we need a combination of techniques.

Only the physical techniques can cost effectively deal with the short path differences needed to achieve better than 2m accuracy.

Path resolution in the receiver is effective only if there is a direct path so is not useful on its own.

Statistical techniques are effective where enough redundant paths are available.

See Table 15 for a summary across all bands.

6.4 Trade-off table

The various techniques discussed need to be matched with the frequencies available for Licence exempt use. The available frequencies are referenced in CEPT-REC 70-03 [1], shown in Table 14. The comparison with multi-path mitigation techniques is shown in Table 15.

Frequency	6.765-6.795	13.553-13.567	26.957-27.283	40.66-40.7	433.05-434.79	434.04-434.79	868-868.6	868.7-869.2	869.3-869.4	869.4-869.65	869.7-870	2400-2483.5
BW	0.03	0.014	0.326	0.04	1.74	0.75	0.6	0.5	0.1	0.25	0.3	83.5
Power	42dBuA/m @10m		10mW	10mW	a)10mW b) 1mW, 13dBm/ 10KHz	10	25mW	25mW	10mW	500mW	5mW	a)10mW b)100mW
Duty cycle	-	-	10%	-	a)10% b)100%	100	1%	0.1%	-	10%	100%	-
Channel spacing	-	-	-	-	-	25KHz	-	-	25KHz	25KHz	-	-
Carrier λ	44m	22m	11m	7.4m	0.7m	0.7m	0.35m	0.35m	0.35m	0.35m	0.35m	0.12m
Frequency difference λ	10Km	21km	920m	7.5Km	170m	400m	500m	600m	3Km	1.2Km	1Km	3.5m

Table 14: Licence exempt bands under CEPT REC 70-03

Frequency band	6.7-40.7MHz	433-434MHz	868-870MHz	2400-2483MHz
Antenna design				
Cross Polarisation	Antennas lose their cross polarisation characteristics as they are scaled down. Therefore appropriate antennas would have very poor polarisation characteristics. Antenna arrays for steering and diversity at either station are too big.	Antenna size and separation at the receivers is still an issue. An antenna on each shoulder would achieve the separation, but the distance would be uncontrollable. Null steering / rotating could be used at the base stations.	Possible, patch antenna approach. Bespoke design required to address size issue.	Antennas of suitable size feasible. Starting to appear on market due to indoor WLANs.
Beam steering			All techniques feasible, at either end of the link.	
Null steering				
Selective diversity				
Combined diversity				
System redundancy				
Base stations	More base stations viable and essential for some bands, as the only mitigation technique.			
Mobile	Multiple antennas not viable due to separation required.	Multiple switched / complex combined antennas possible.		
Frequency spreading				
Direct Sequence. Based on MSK null-null BW	Bands are too narrow to counter multipath. Slow FHSS and base station redundancy could be used against interference.	129m path difference separable	320m path difference separable	2.7m path difference separable
Frequency hopping		Counter interference only		
Frequency sweep		As for DSSS		
Channel equalisation	Same arguments as for DSSS.			

Table 15: Trade off table for multi-path mitigation in licence exempt bands

7 SYSTEM REDUNDANCY ISSUE

With a deployment of multiple antennas around a village it is possible to increase the raw accuracy.

Each measurement from each base station has an uncertainty associated with it, σ. If we assume that every measurement made at a particular location has the same uncertainty (they are all equally affected by multipath and noise effects) then multiple measurements will result in a lower uncertainty. Therefore, increasing the number of line of sight transmitters seen at every location will have a beneficial effect.

To model this affect we can assume that the *n* deployed transmitters are located around the village at an equal spacing. This means that they are distributed in *n* sided polygons around the outside of the village. Every location within the village can now be checked for the accuracy achieved from a purely geometrical view. The minimum requirement to achieve a two dimensional fix for a particular location is three and so the smallest polygon considered is an equilateral triangle. Figure 14 shows the average error around the village for each number of base stations seen. It is noted that these are direct line of sight base stations and so if these can be identified as being direct line of sight the total number of transmitters seen at each location may be greater than the total number of line of sight transmitters.

As can be seen in Figure 14 the average fix improves with the number of deployed transmitters (or with the number of direct line of sight signals detected) as expected. However, the increase in accuracy is not significant for the increase in cost of deploying a large number of transmitters when more than about 10 are deployed (or seen by each location). The percentage error shown in this Figure is relative to σ (the error from one transmitter). Therefore, measurements of line of sight from 10 transmitters will lead to better than half of the theoretical error due to one transmitter.

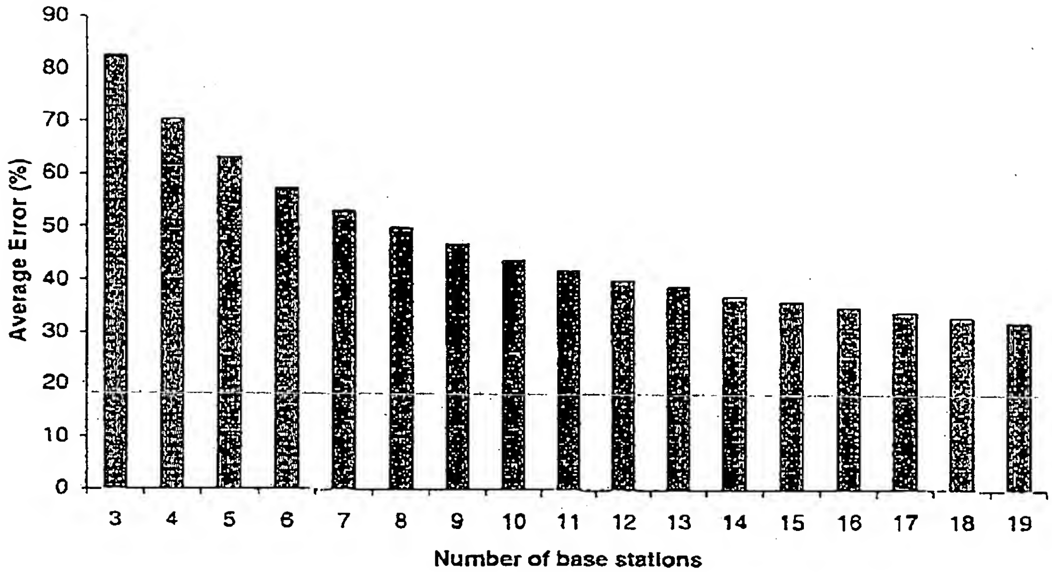


Figure 14: Average error calculated for location calculations around a village as a percentage of the theoretical error from a single measurement. For example, line of sight measurements for 8 base stations results in an improvement in the error by a factor of 2.

8 MULTIPATH PROPAGATION MODEL

This document describes a 13 ray multipath propagation model that has been developed to study the multipath environment within rectangular rooms and similar walled structures. This model has been used to investigate the effects multipath has on phase based location systems and the improvement various multipath mitigation techniques provide. Results for two scenarios, a room and a closed-off urban street are presented.

8.1 Model description

8.1.1 Outline

The model is based on calculating the pathlengths between a transmitter and receiver for the direct, first bounce and second bounce multipath components in a 2D box. With the reflection attenuation experienced with imperfect reflectors, the power of further components are considered to be too small to warrant modelling. If they were modelled, other inaccuracies in the input data and the model detail would probably limit their worth anyway so their not considered. Environments with perfect reflectors (metal walls) would require the model to be extended, to maintain accuracy.

Along with the pathlengths, the angles of reflection and the relative permittivity of the wall material, are used to calculate the following for each component arriving at the receiver:

- Horizontally polarised power,
- Vertically polarised power,
- Delay,
- Phase shift.

Relative (to the direct path) figures are then given below. This information is used to calculate the actual total multipath component present in the channel, which can be converted to SNR, Phase error and ultimately location error.

8.1.2 Input details

The model requires the following inputs, referring to Figure 15.

- Maximum room size, Xrm,Yrm
- Position of receiver (fixed), Xr,Yr
- Maximum position of transmitter Xtmax, Ytmax
- Minimum position of transmitter, Xtmin,Ytmin
- Number of transmitter positions from Xtmin to Xtmax,

- Number of transmitter positions from Ytmin to Ytmax,
- Wavelength of signal, λ metres
- Relative permittivity of wall material, ε_r

As is seen, the receiver is fixed and the transmitter can be moved around the room to any number of positions (software limits to 25,500 points per simulation), within user specified bounds to collect statistical and locations specific information.

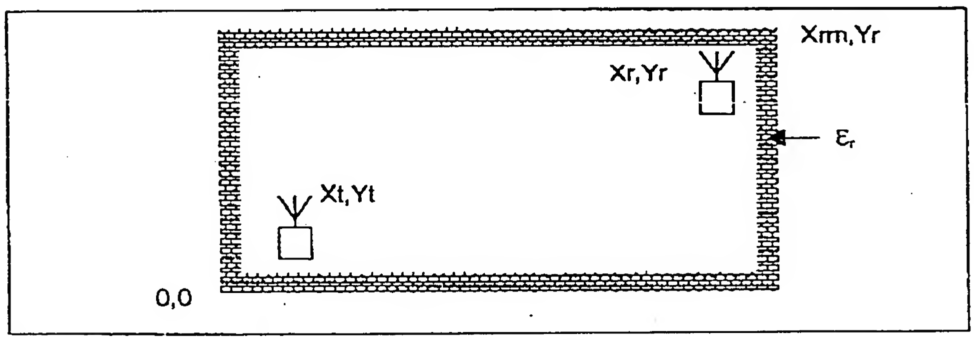


Figure 15: Model dimension inputs

8.1.3 Paths and reflections

The model calculates the following paths, referring to Figure 16.

- 1 direct path
- 4 single bounce paths
- 8 double bounce paths

The reflection angle used is also shown in Figure 16 and defined as the angle between the incident wave and the plane of reflection, (the wall). Note this is different to the traditional optical physics definition.

The reflection coefficient equation depends on the polarisation of the incident wave and is usually referred as the incident E field being either *perpendicular* or *parallel* to the plane of incidence. The plane of incidence is the plane containing both incident and reflected rays. Thus the definitions are independent of antenna polarisations with respect to the ground / wall or any other physical reference plane. As an example, a signal from vertically polarised monopole reflecting horizontally off a wall next to it, would have its E field perpendicular to the plane of incidence and therefore the perpendicular reflection coefficient would be used. For the same antenna reflecting off the ground (as in the traditional 2-ray propagation model) the E field would be parallel to the plane of incidence and hence the parallel coefficient equation would be used. The reflection coefficients for concrete at 2.4GHz (εr=3.5) and 400MHz (εr=10) are shown in Figure 17.

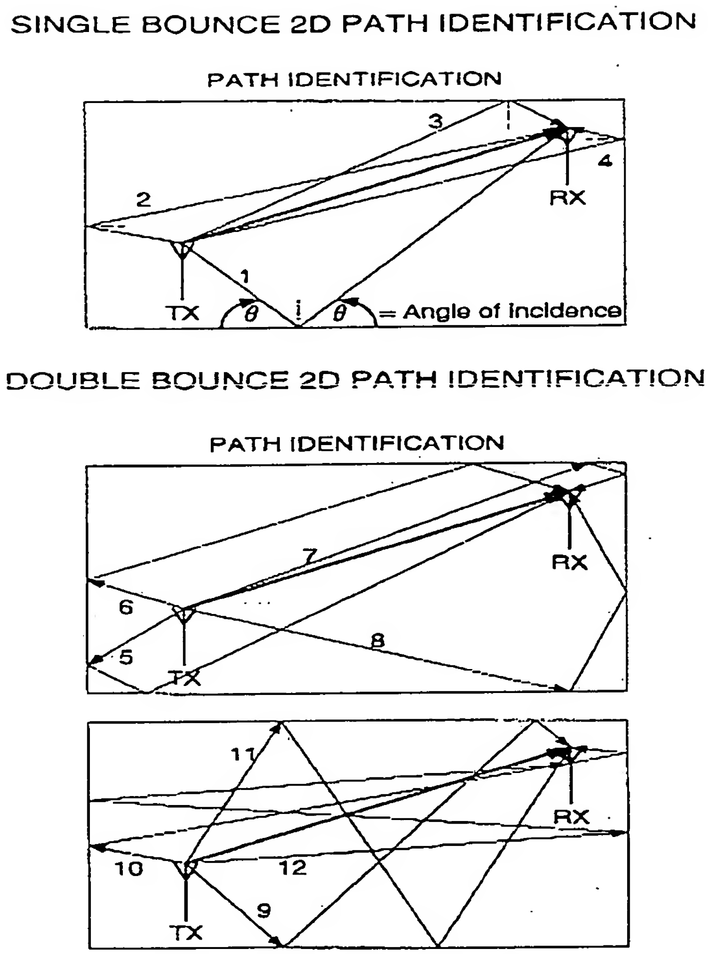


Figure 16: Multipath component identification & incident angle definition

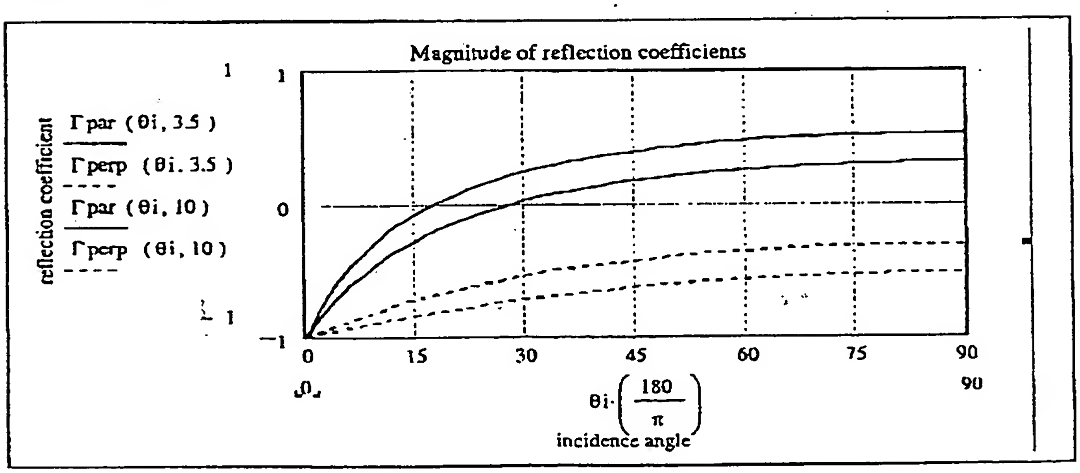


Figure 17: Reflection coefficient vs incident angle

As is seen, the perpendicular component always has a phase inversion, whilst the parallel component is only inverted below the Brewster angle, the angle when $\Gamma = 0$. Note also, the higher the relative permittivity, the lower the Brewster angle and the larger the magnitude of the reflection coefficient. From this and other material data, several conclusions can be drawn.

1. The lower the frequency the higher the permittivity and the stronger the reflected power.
2. Circular polarised waves change "hand" for a reflection greater than the Brewster angle.

The first might seem odd since power transferred through wall materials also increases as frequency drops, but the power absorbed by the dielectric also drops, so the conservation of energy is preserved.

The second indicates a possible multipath mitigation benefit as single bounce reflections would be attenuated by the cross-polarisation ration of the receiver antenna. A link between a pair of right-hand circular polarised antennas would reject the first bounce component as it would be left hand circular polarised. Unfortunately, as the reflected signals go through more reflections and the horizontal and vertical power components are attenuated by different amounts, the unity axial ratio is no-longer preserved and the receiver antenna looses its resilience to the multipath components. Hence circular polarisation can only generally be considered effective against single bounce components.

The model only considers specular reflection, as described above. Diffuse scattering of the signals, caused by rough surfaces is not considered as this only becomes an important factor when the Rayleigh Criterion is reached.

ie the height of the surface roughness is more than:

$$h_{critical} = \lambda / 8\cos\theta_{incid}$$

For the frequencies of likely interest, the reflectors will not be rough enough. Even if this was encountered, the specular model already indicates the worst case multipath limit.

8.1.4 Multipath power

As the multipath components are not completely separated at a basic receiver input, the magnitudes vector add to produce a resultant interfering power at that instant in time. The *Average* power would simply be the addition of the average powers for each component, if they could be isolated (as in a Rake receiver). The total instantaneous multipath power is then:

$$|r(t)|^2 = \left|\sum_i a_i \exp(j.\theta_i(t))\right|^2$$

where a_i and $\theta_i(t)$ are the amplitude and phase of the i'th multipath component.

Note, as this is an instantaneous power, for the particular location and time, it is quite possible to get $|r(t)|^2$ values up to twice the average power received. Averaging $|r(t)|^2$ over time would result in the same average power.

8.1.5 Signal to Noise Ratio and Phase error

Figure 18 shows how the total multipath could add to the wanted direct path signal to give a resultant A_r. The SNR, describing the ratio of signal power to multipath "noise" power is defined in the model as $1/a_{tot}$ as the powers are relative. The SNR does not take account of the phase, to highlight the possibility of zero multipath phase but strong multipath power: pure constructive of destructive interference. The phase error θ_r, is given by:

$\theta_r = \tan^{-1} (\sin(\theta_{tot}))$ noting again the wanted LOS signal is unity.

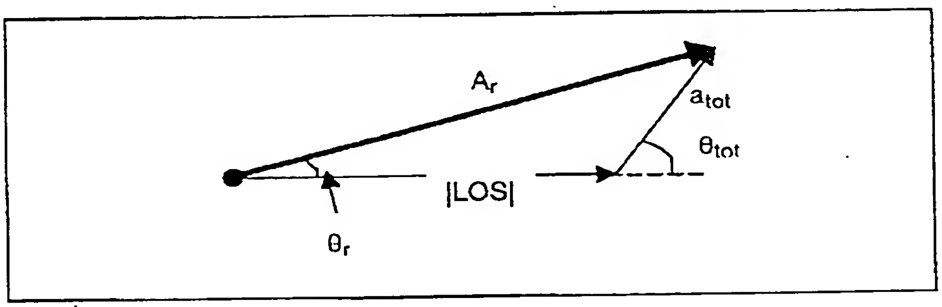


Figure 18: Multipath adding to wanted LOS path

8.2 Accuracy for phase based location systems

The location systems under investigation measure the phase of incoming signals and convert that to a fraction of the wavelength to get a distance measurement. See the system design section for more details. Any errors in phase produced by thermal noise, phase noise, filter imperfections or multipath lead directly to location errors. The first three can be managed through careful deployment layout to keep within range and quality electronic design. The last needs to be dealt with separately, using the techniques discussed in Section 6. A resultant multipath component will introduce a phase error θ_r. This, as a percentage of 360^o, will provide the distance error of the wavelength.

For example, with a phase difference system with 15MHz separation between tones, equating to a 20m wavelength, a 7.2^orms phase error would give 2% on 20m or 0.4m rms distance error. A 7.5MHz tone separation could only tolerate 3.6^o phase error to give the same physical distance error. Hence to address multipath errors, either the multipath has to be addressed directly, or the percentage error must be accommodated in the frequency selection to achieve the required location accuracy.

SNR is often easier to utilise, the SNR representing worst case phase error (multipath at 90^0 to the wanted LOS component) verses rms location error is shown in Figure 19. Also shown is the improvement made by averaging samples. Whilst this works for random noise, it will not be effective against multipath unless that changes sufficiently during the sample period.

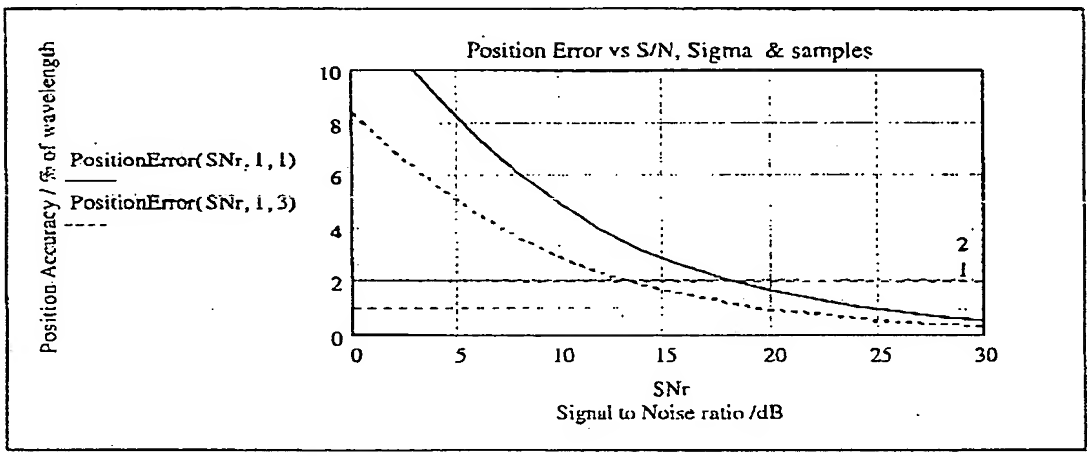


Figure 19: Phase based location system error verses SNR

The model uses the calculated phase error due to multipath to calculate the position error, with respect to the wavelength of the signal. No other sources of phase error are considered.

8.3 Multipath mitigation modelling

Several multipath mitigation techniques have been very simply modelled, to provide an indication of their effectiveness. Although more complex modelling is quite possible, it is considered that the chosen methods are sufficient for comparison purposes.

8.3.1 Direct sequence spread spectrum

In a DSSS receiver the ability to separate multipath components is limited by the chip period Tc. Strictly speaking, components less than 2Tc apart would not be separated due to the correlator output being spread over 2 chip periods, however with a typical Gaussian spread of components Tc seconds is considered the minimum separable difference. Whilst n separately identified components could be added coherently to improve SNR by $\sqrt{n}$, as in a rake receiver, for this model, a simpler architecture is used, where only the components within the first chip period Tc are considered in the SNR and phase error calculations; all others are ignored.

8.3.2 Circular polarisation

As for the first reflection, the power of the two components of the wave are generally still closely matched, reasonable cross-polar rejection is expected at the receive antenna. Therefore, without adding in antenna specifications etc, 10dB of attenuation is added to the single bounce paths. This is considered a realistic level given typical antenna specifications and imperfections etc.

Note that for electrically very small antennas, circular polarisation is rarely properly formed so the effectiveness of this technique would be limited. The model assumes proper circular polarised antennas are being used.

8.3.3 Combination of both

A third model combines both the above techniques to illustrate the effectiveness of the combination.

8.4 Example results

8.4.1 In room scenario

A room described in Table 16 was analysed. Figure 20 shows the statistics of the delay and vertically polarised power information from moving the mobile around the room. Figure 21 shows the resulting basic location error, whilst Figure 22, Figure 23 and Figure 24 show the improvement made from a DSSS system operating with 22ns chip periods, circular polarisation, and a combined system.

	X	Y
Room -rm	10	5
Receiver -r	9.9	4.9
Transmitter -tmin	0.2	0.3
Transmitter -tmax	9.7	4.6
Number of Tx positions	100	50
Wavelength λ	0.125	
Wall permittivity εr	3.5	

Table 16: Room scenario, information

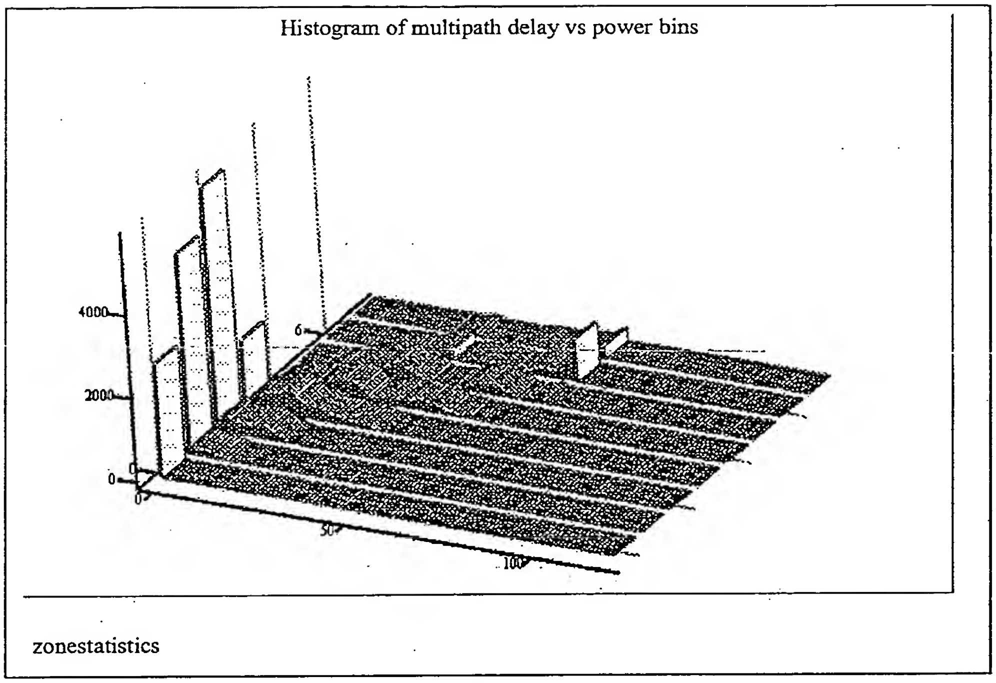


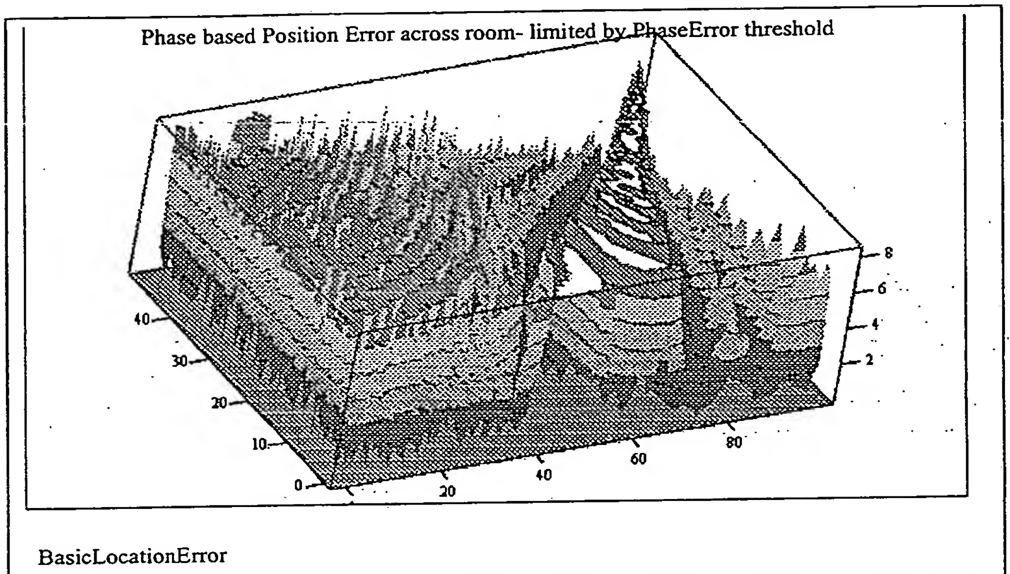
Figure 20: Room multipath power/delay histogram.
X axis = 1ns delay steps, Y axis = 5dB power steps, Z axis = number of paths.

Figure 20 shows that for the presented positions, most of the multipath is very close in and within 20dB of the LOS power. This is due to reflections coming from the walls behind the base station receive antenna. These would be removed by using reflector style antennas pointing only into the room. Similar reflections occur when the mobile is itself close to a wall. These could be removed by having multiple sectored antennas and switching between them. There is a second large peak around 66ns, albeit at lower power, which is due to the double bounce signals and the long dimension of the room. In this dimension, the pathlengths change very little with the Y position due to the fine angles involved. Double bounce signals from the short dimension are around the 30ns period and are more smeared by the axial ratio of the room increasing the angle changes with position and increasing the range of pathlengths. The power level grouping of these signals indicate the different pathlengths and reflection coefficients covered.

Figure 22 to Figure 24, showing rms location accuracy relative to the measured wavelength, are scaled with the number of sample locations across the room defined in Table 16. Errors above 8% are limited to 8% (shown as red). For the 15MHz recommended fine resolution tone separation location errors up to 5% (half way across the green contour) can be tolerated for 1m accuracy, whilst 2.5% gives 0.5m accuracy (dark blue & purple). For this scale of zone, it is seen that DSSS at 22ns chip period offers little protection, only smoothing off the errors

whilst most improvement is gained from circular polarised antennas. The mean and standard deviation figures for each plot describe the spread of location error across the room.

Figure 23 shows that the vast majority of the room should be covered with sufficient accuracy to meet the +/-0.5m rms requirements and certainly to +/-1m rms.



Mean(BasicLocationError)=3.684   StdDev' (BasicLocationError)=2.864

Figure 21: Basic location error / % of λ

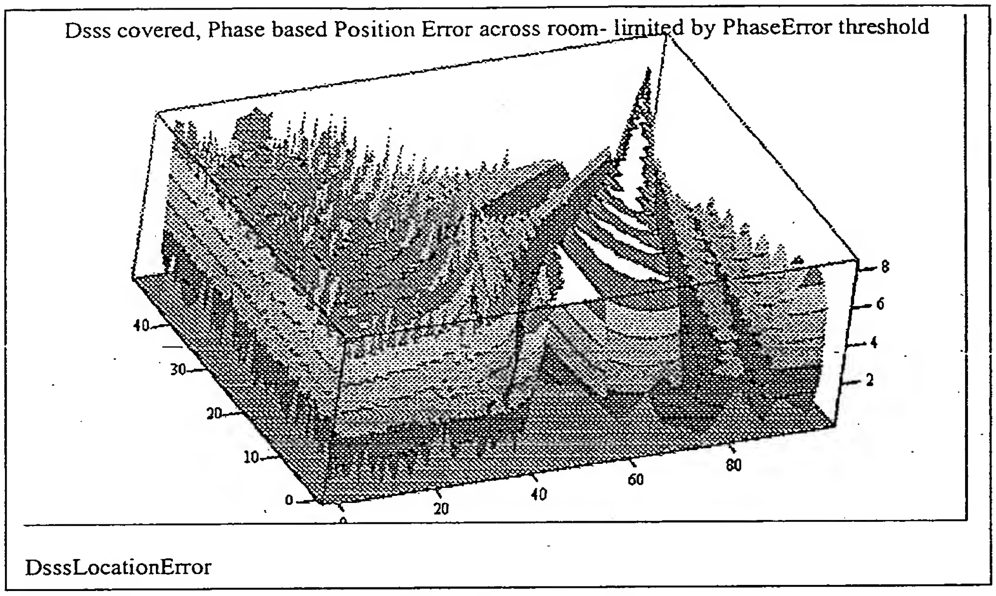
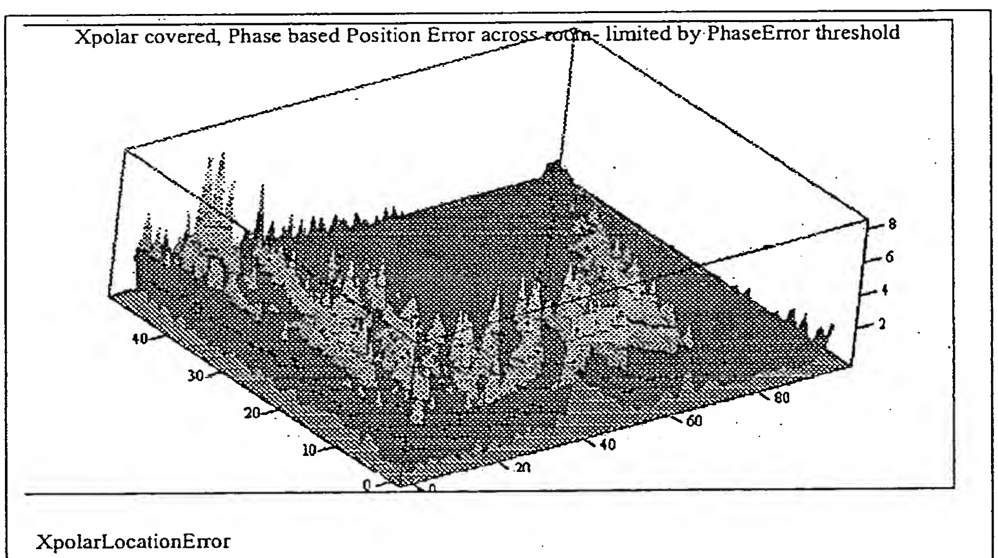
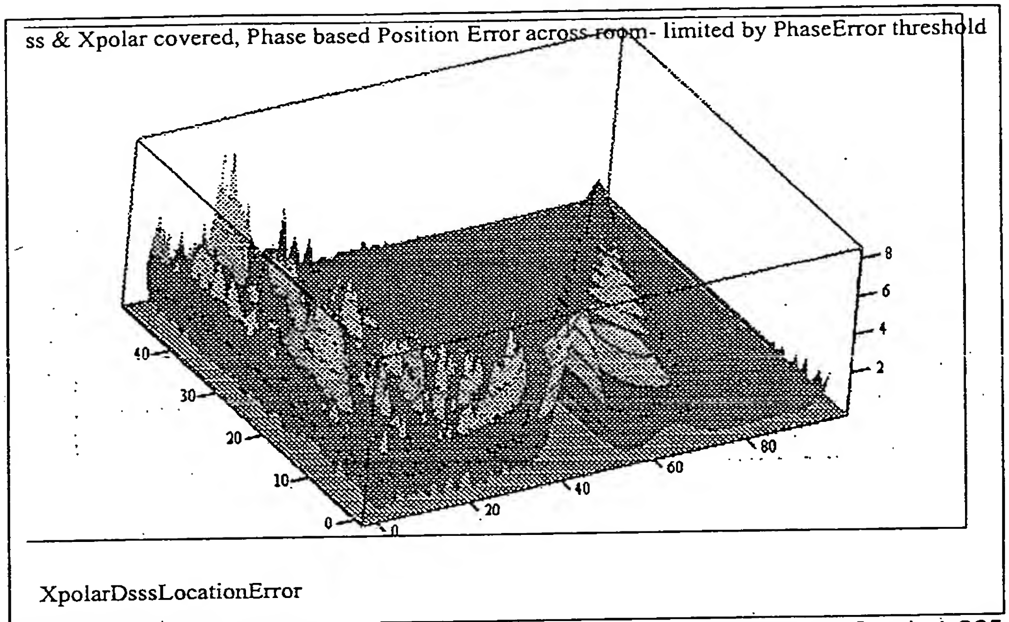


Figure 22: Location error with DSSS, 22ns chip period
Mean(DsssLocationError)=3.645 StdDev' (DsssLocationError)=2.838



Mean(XpolarLocationError)=1.061 StdDev' (XpolarLocationError)=1.308

Figure 23: Location error with circular polarisation



Mean(XpolarDsssLocationError)=0.978 StdDev'(XpolarDsssLocationError)=1.225

Figure 24: Location error with circular polarisation and DSSS

8.4.2 In street scenario

A closed ended street described in Table 17 was analysed. Figure 25shows the statistics of the delay and vertically polarised power information from moving the mobile around the room. Figure 26 shows the resulting basic location error, whilst Figure 27 to Figure 29 show the improvement made from a DSSS system operating with 22ns chip periods, circular polarisation, and a combined system.

	X	Y
Room -rm	200	15
Receiver -r	150	14.9
Transmitter -tmin	0.5	0.2
Transmitter -tmax	199	14.6
Number of Tx positions	200	100
Wavelength λ	0.125	
Wall permittivity εr	3.5	

Table 17: Street scenario, information

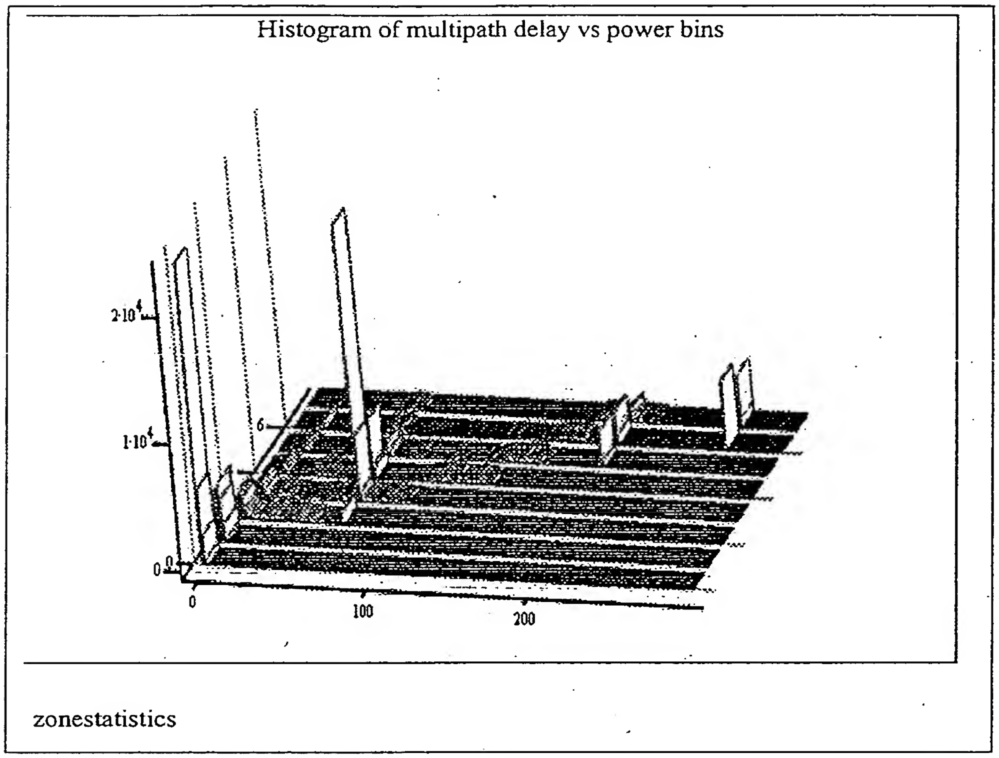


Figure 25: Street multipath power/delay histogram.
X axis = 1ns delay steps, Y axis = 5dB power steps, Z axis = number of paths.

As with the indoor system, the peaks with very short excess delays are due to either of the stations being very close to a reflecting wall. The next main peaks around 300ns, 1000ns and 1350ns are due to the relationship between the selected position for the base station and the zone dimensions. For a 200m long zone with a fixed base station 50m from one end, these excess delays correspond to excess pathlengths 100m, 300m and 400m long. The aspect ratio of the zone and the resulting fine angles reduce the spreading around these peaks to within a few delay bins.

The location accuracy plots of Figure 26 to Figure 29 show that there is sufficient multipath to render an unprotected implementation useless everywhere but extremely close to the base station. However, the use of circular polarised antennas dramatically improves the situation. DSSS this time does show a noticeable improvement but it is still secondary. The reason one end of the street is still poor is that these areas are mostly affected by double bounce components which the antennas don't affect and the excess delays from corner bounces are still too short for the DSSS to be effective. In reality, since such a street would be covered by several base stations, this would not be a problem and the location information from this base station would be made redundant within the location algorithm process

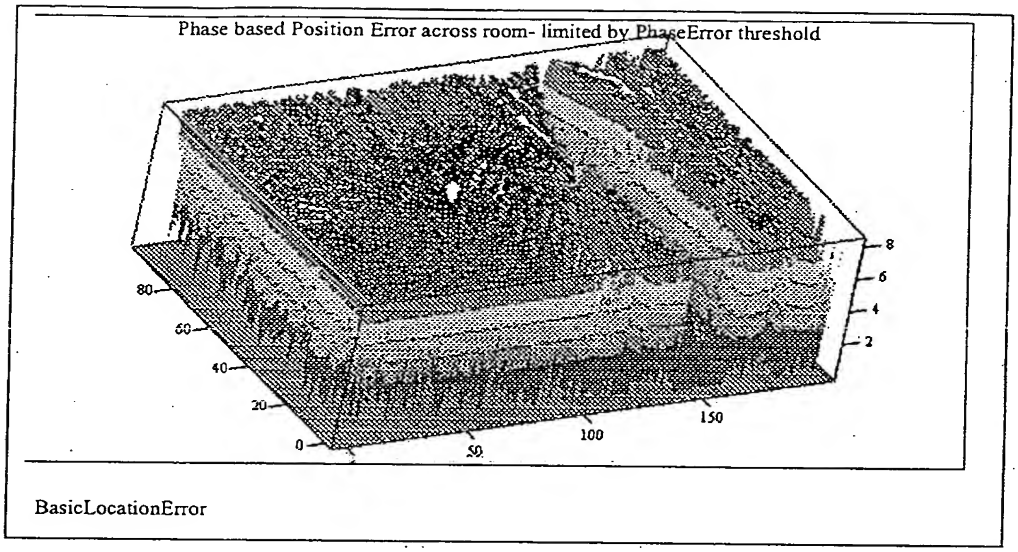


Figure 26: Basic location error / % of λ
Mean(BasicLocationError)=5.296   StdDev' (BasicLocationError)=3.141

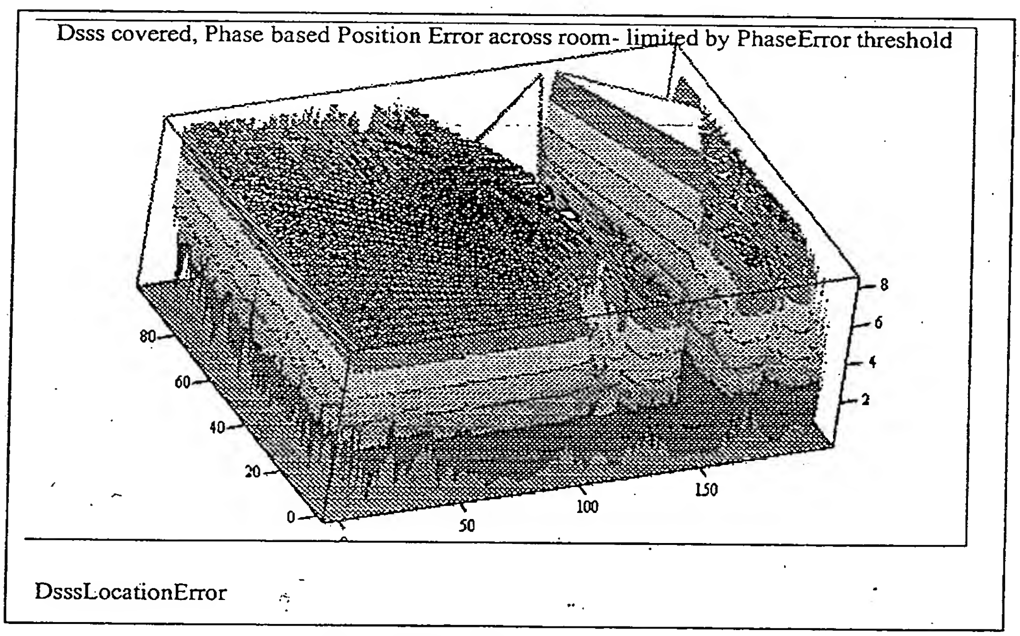


Figure 27: Location error with DSSS, 22ns chip period
Mean(DsssLocationError)=5.256   StdDev' (DsssLocationError)=3.076

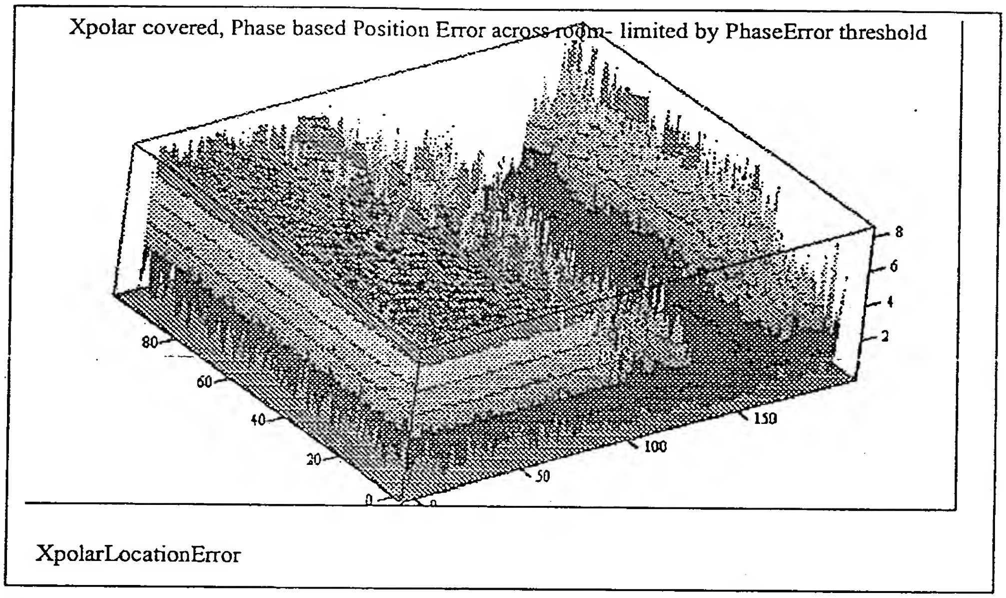


Figure 28: Location error with circular polarisation.
Mean(XpolarLocationError)=2.664 StdDev' (XpolarLocationError)=2.840

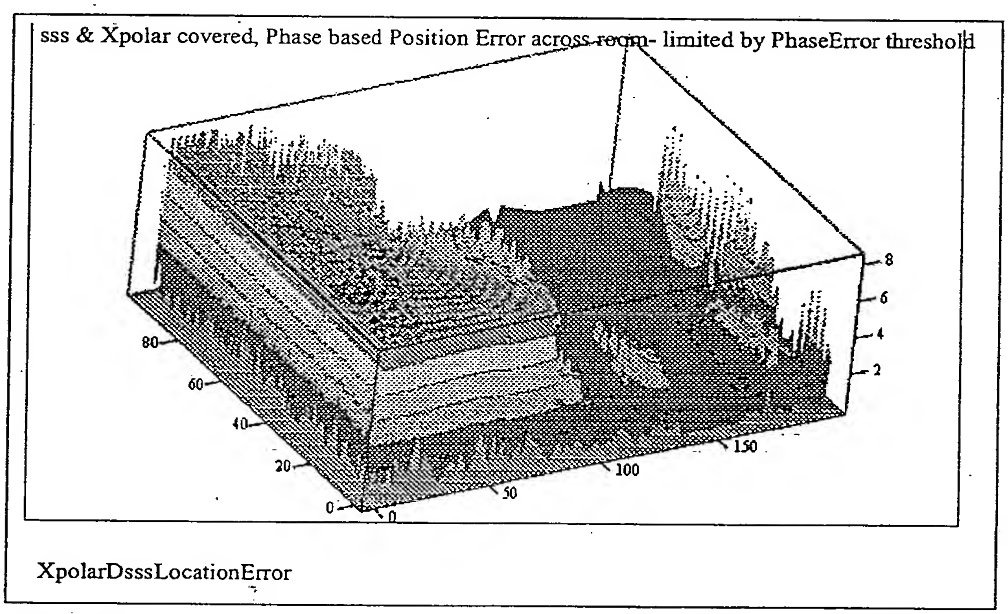


Figure 29: Location error with circular polarisation and DSSS
Mean(XpolarDsssLocationError)=2.152 StdDev' (XpolarDsssLocationError)=2.57